# Proceedings of the I·R·E

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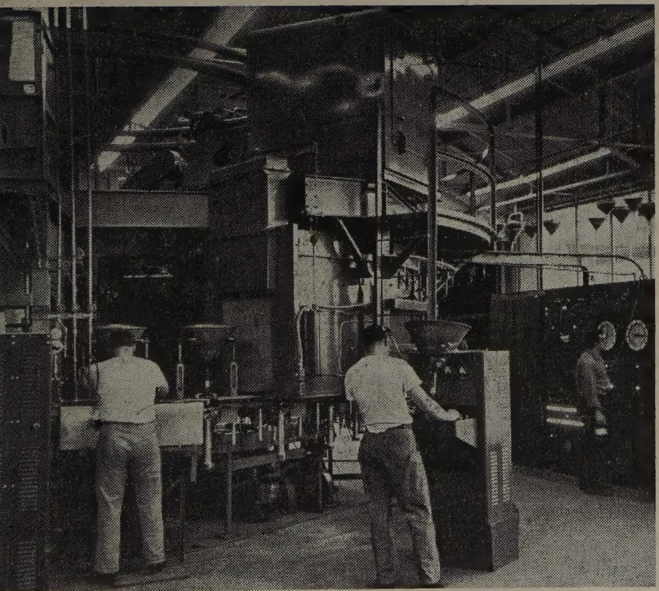

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IRE Standards on Pulses
Transistor Trigger Circuits
Microphone Sensitivity Measurement
Trigger-Grid Thyratron
Tropospheric Scattering of Radio Waves
Radio Spectra Calculation
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VHF-UHF Tail-Cap Antenna
Patterns of Antennas Near Elliptical Cylinders
Current Distributions on Helical Antennas
Atmospherics Waveform Receiver (Abstract)
Directive Antenna Response to Incoherent Radiation
Optimum Operation of Echo Boxes
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Linear Varying-Parameter Networks
Distortion in Linear Passive Networks
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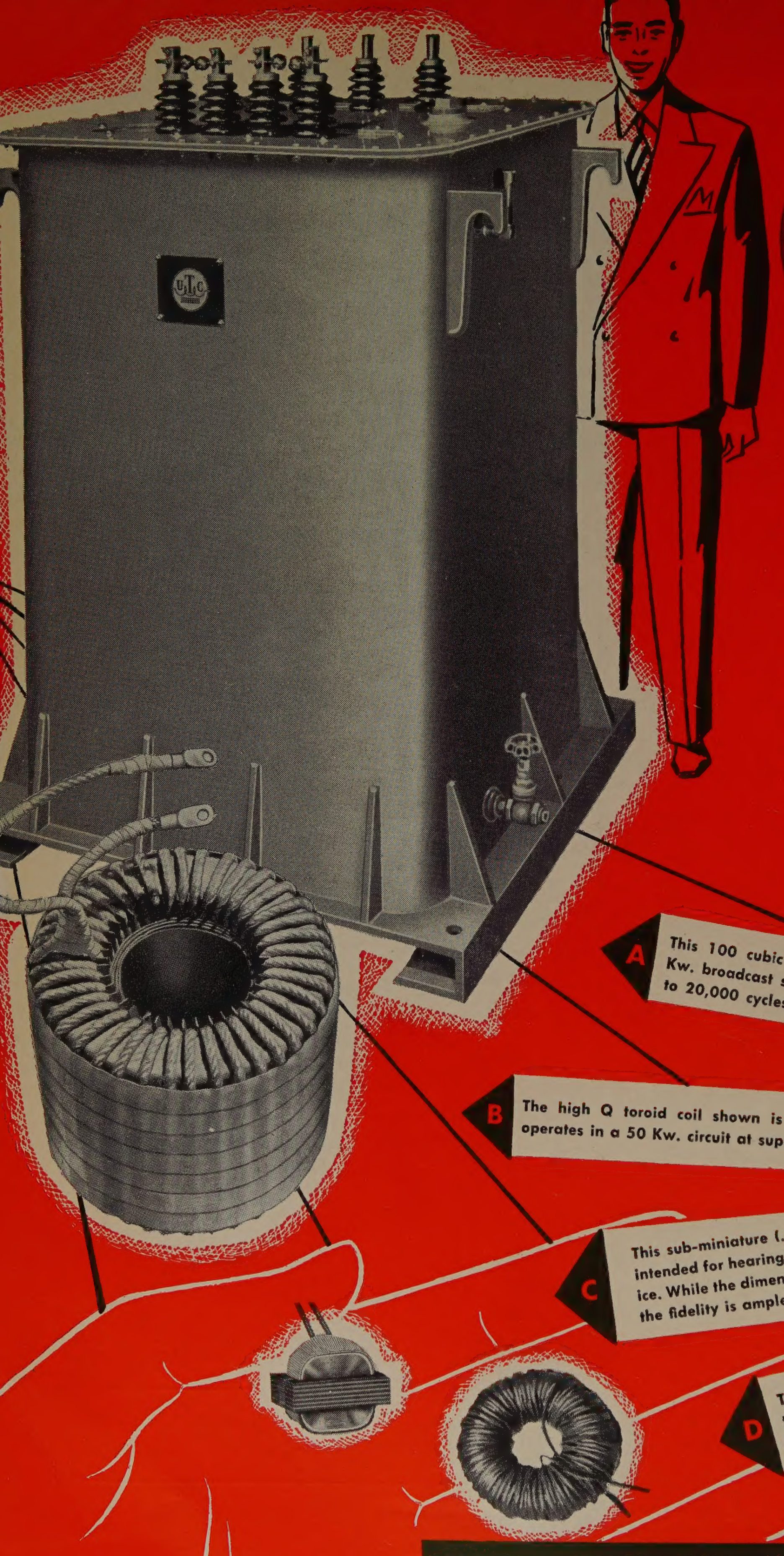
The Institute of Radio Engineers



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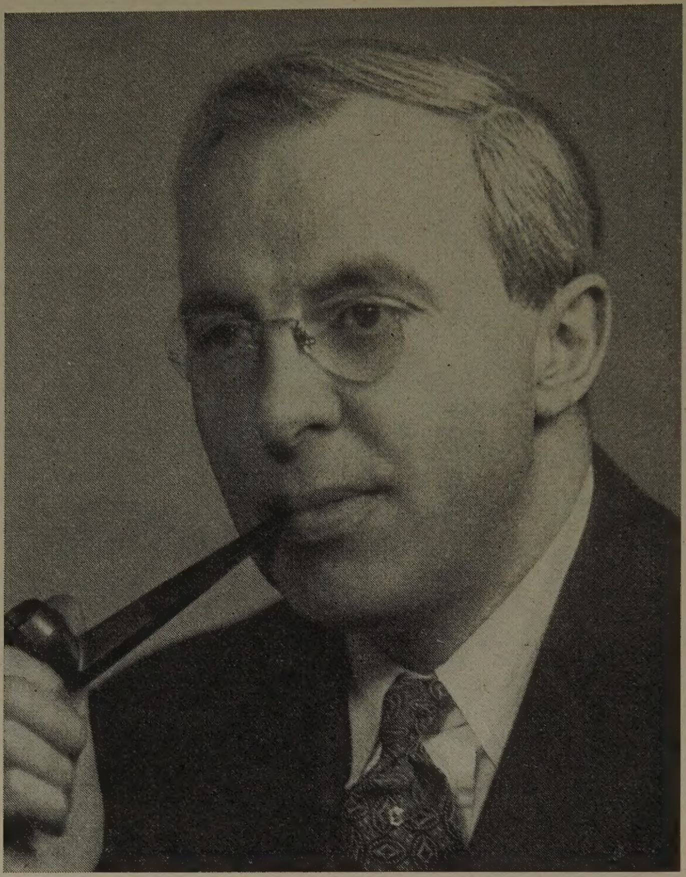

# Allan B. Oxley

Regional Director, 1951-1952

Allan B. Oxley, chief engineer of the home instrument division, RCA Victor Company, Ltd., Montreal, was born in London, England, in 1901. After attending night school for 13 years, he took a postgraduate physics and atomic theory course at the University of Toronto in 1929.

He began his career as an apprentice instrument maker, and became a full-fledged instrument maker in 1922; at that time he engaged in the manufacture of microscopes, spectrometers, and fine scientific instruments. From 1924 to 1926 he was mechanician to the department of physiology, University of Toronto, and assisted Sir Frederick Banting in the development of insulin for diabetics. Later, when the Oxley and Meredith Company was created, Mr. Oxley was also engaged in the organization of the IRE Canadian Section. In 1927 he was appointed manager and chief engineer of the King Radio Corporation, and in 1929 became chief engineer of the Williams Piano Company, which had absorbed the King Corporation. In 1933, as chief engineer of the Philco Radio and Television Corporation, he was sent to England to organize a London factory. He returned to Canada in 1935 to become, first, quality engineer, and later, chief engineer for the RCA Victor Company, Ltd., in Montreal.

During World War II he was responsible for a number of important electronic research developments for all branches of the British, Canadian, and U. S. governments.

An Associate of the Institute in 1925, Mr. Oxley assisted in the organization of the Montreal Section. He became a Member in 1933, and a Senior Member in 1946. In 1933 he served as Chairman of the Toronto Section, in 1939 and 1949 as Chairman of the Montreal Section, and is now Regional Director of the Canadian Region.

Mr. Oxley was chairman, also, of the Engineering Committee of the Radio Manufacturers Association of Canada, in 1939, 1940, 1941, 1944, 1945, 1946, and 1949. He was general co-ordinator of the Canadian Radio Technical Planning Board from 1944 to 1948, and has been Canadian representative on the United States RTMA Receiver Committee since 1939. He is a member of the Royal Canadian Institute.

# Why Awards

ERNST WEBER

In an age when the long-standing antagonism between far-visioned idealism and allegedly "practical" materialism is especially intense, the following address becomes a particularly significant and encouraging "confession of faith." Its author is director of the Microwave Research Institute, and head of the electrical engineering department at the Polytechnic Institute of Brooklyn, and is also Chairman of the Subcommittee on Tutorial Papers of the IRE Education Committee.—*The Editor.*

One of the highlights at the Banquet during the 1951 IRE National Convention was the bestowal of five specific prizes and awards upon members of the Institute who have contributed in an outstanding manner to the advancement in the art and science of radio and electronics. Three of these awards are in themselves memorials to outstanding late members of the Institute, Harry Diamond, Morris N. Liebmann, and Browder J. Thompson; the fourth is the Editor's award, and the fifth, the highest award, is the Medal of Honor.

In addition, the Institute confers the honorary Fellow grade upon a varying number of members who have distinguished themselves both professionally and through service to the Institute.

Having witnessed several of these memorable occasions and having been asked to respond for the newly elected Fellows at the most recent Convention, I began to reflect seriously about the basic reasons for this tradition. What is at the root of these honorary awards, and what is it that the Institute obtains in return?

Surely all of us have wondered from time to time just what makes an organization grow and expand. Or perhaps, since organizations are composed of people, what motivates a person to belong to a group, to join in an effort transcending his own realm? There are some cynics who pride themselves with superior intelligence, and who hold the view that man does *only* what benefits him as an individual or what gives him personal pleasure—and that there is no such thing as caring for others, or working because of a sense of obligation to mankind. At least as a teacher, I must object most strenuously to such a viewpoint—it is contrary to all basic tenets of educational philosophy! Such a viewpoint is as ruinous for the family as it is for the nation. We certainly would never have this Institute of Radio Engineers if personal motivation were as empty and barren as these cynics would have us believe.

Any social institution *is* what its members make it! If its members do not care, if they let "the others" worry, let "the others" be the fools to work—then that institution, however great it might appear at the moment, is *doomed;* it will crumble and disappear. If on the other hand, there are enough members who *actively* support the institution, who freely devote time and effort to its tasks and obligations, who share in the belief of a common goal and of common interests in higher achievements, then that institution will grow strong and prosper.

Looking back at the growth of The Institute of Radio Engineers, one cannot but conclude that it must have had a strong core of devoted men who made it what it is today, who gave, and still are giving unstintingly of their time and effort for the promotion of its services to the members. These men have created the meaning of honors and awards conferred by the Institute, they have made the "Fellowship" in this great professional Institute worthy of its distinguished connotation!

I take it, then, that awards are the tangible expression of encouragement to continue the individual professional contributions with renewed and greater effort in order to serve better this Institute, our profession, and thus, our country.

# What the Engineers Have Done to Television*

ROBERT C. SPRAGUE†

MY APPEARANCE here tonight reminds me or an experience I had when learning to ski a few years ago. On looking down the precipitous slope I asked myself, "What am I doing here?" I might ask myself a similar question tonight. "What is a parts manufacturer doing addressing a gathering of engineers?"

Of course I realize that the invitation to speak here resulted from the fact that in a weak moment last June I accepted the job of president and chairman of the board of the Radio-Television Manufacturers Association. Fortunately, I will soon unload the responsibilities of presidency on the shoulders of another and younger man.

I do not need to tell you that the job of heading an Association which represents an industry as volatile, dynamic, and explosive as the radio-television industry is not a moonlight cruise on calm seas. The hazards of accepting such a responsibility, like those of public speaking or letter writing—the latter of which have been recently illustrated at the White House—are many and the rewards, if any, are uncertain.

However, I would be less than frank if I did not admit that there are many compensations, and most of us in this situation are like the farmer who said that he was going to town Saturday night to get drunk and complained, "Oh how I hate it."

I do feel at home, however, in speaking to such a representative group of our industry's engineers although for some years I have lived on the management's side of the railroad tracks. While there may be a wide difference of opinion as to whether management, sales, and promotion or technological developments are chiefly responsible for the remarkable development of radio and television in the last quarter of a century, I note that the radio engineers were at least smart enough to organize first. I refer to the fact that The Institute of Radio Engineers was founded in 1912, while RTMA was not formed until 1924.

Seriously, I believe that the radio-television industry's achievements may be credited to three major factors: research and development, production efficiency, and alert management and salesmanship. Together they have made possible the mass production of high-quality radio and television receivers, as well as other electronic equipment so important to our Armed Forces and to other industries.

The subject I had adopted tonight "What the Engineers Have Done to Television," may sound facetious. I assure you that it is not. It is obvious that were it not for the engineers, there would be no television. Less widely known is the fact that there would be no television as we know it today except for the unselfish work of a large number of engineers who have consistently sought to protect the public interest, as well as the interests of the industry, by resisting experiments conducted at the expense of the consumer's pocketbook. It is certain that if the action of these engineers, and of the industry generally, had been only self-seeking, or faulty, or erroneous—as has been recently implied by a member of the Federal Communications Commission—that the public would have rebelled and television would not be the multibillion dollar industry it is today.

In a supplemental opinion issued by Commissioner Robert F. Jones in connection with the FCC's preliminary findings in the color television case, the "good faith, truth, and veracity" of prominent industry engineers were questioned. Commissioner Jones stated that the hearing had made it "crystal clear that the industry's engineers were unsound analysts," and in 1946 and 1947 had given testimony which was "completely worthless" because "their economic interests blinded their engineering judgment." He added, moreover, that there was "grave doubt" that these engineers could "be relied upon to predict the potential performance of any system whose adoption might prejudice their economic interest." The Commissioner also charged that the industry attempted to "confuse the Commission" by "sham engineering testimony."

A judicial examination of these charges made by a responsible government official raises the following interesting questions affecting professional engineers and their relationships to their industrial employers:

Are honest differences of engineering opinion legitimate?

Do engineers understand more fully systems and equipment at the later stages of development than in earlier stages?

Accordingly, may earlier, and honest engineering opinions, require subsequent and legitimate modification?

Is it justifiable for professional engineers (as it is for medical doctors) to require detailed technical knowledge and extensive tests (clinical experience) before approving and widely adopting something new?

Should economic factors (that is, the cost of new and replacement apparatus, obsolescence of older apparatus, high manufacturing costs resulting from changeovers) be considered by engineers in determining the value of a system or equipment? If not, how should engineers be "insulated" in an "economic vacuum?"

In general, are engineers less sound analysts than statesmen, politicians, other professional men, industrialists, and so on? If they are, what is the reason for their inadequacy?

Are engineers more prone to offer fallacious, sham, questionable, or worthless conclusions than any other of the previously mentioned groups? If so, why?

I do not think it necessary for me or any one else to come to the defense of our industry's engineers. Their record for more than a quarter of a century speaks for itself. However, I would like to take this opportunity to deny categorically the charges which have been hurled at our industry and to comment briefly on what may properly be expected from the engineering fraternity.

Engineers are highly individualistic persons who often do not agree with each other or their commercial colleagues. They are influenced in their decisions by facts and figures, and not by pressure from whatever source.

It is doubtless true that engineers recognize economic factors and hesitate to spend money recklessly, whether it comes from the ultimate consumer or their employer, and I admit that this is rather a novel characteristic in these times.

Our industry engineers generally strive to offer to the public the fruits of their technological developments at the lowest possible cost, but they seek to avoid foisting on the public either systems or equipment which have not been thoroughly tested.

It is true that engineers are sometimes brutally frank in expressing opinions, whether they speak as individuals or representatives of industry or engineering groups, but I think that any fair-minded person would concede that they are honest and fair in their ultimate judgments.

Perhaps a brief review of the early history of television and the role played by engineers in its development will be timely. After early experiments, black-and-white television had developed to a point early in 1940 where the Radio Corporation of America made plans to market television receivers and to place commercial television stations on the air. The Federal Communications Commission at that time regarded such action as tending to freeze television standards and thus block its development and discourage further research and experimentation. The Commission in its ruling stated that there should be no commercial TV broadcasting until, and I quote, "the probabilities of basic research have been fairly explored."

The FCC added that as soon as the majority of engineers in industry were prepared to approve any one of the competing systems of broadcasting as standard, the Commission would authorize commercial television operations.

As a result of a meeting between FCC Commissioner James Fly and W. R. G. Baker, director of engineering for many years for RMA and RTMA, and a former President of IRE, the first National Television System Committee was formed. The monumental work of that committee is a matter of record. In six months this committee developed standards and made a report to the FCC. Its technical panels of experts produced reports and minutes totaling 600,000 words, devoted 4,000 man hours to meetings and an equal time to travel, and witnessed 25 technical demonstrations. As you know, the standards for black-and-white television proposed by these industry engineers were adopted and television was off to a successful start, only to be delayed several years by our entry into World War II.

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† Radio-Television Manufacturers Association, Washington, D. C., and Sprague Electric Company, North Adams, Mass.

Perhaps the best measure of success of this almost unparalleled example of co-operation among industry engineers is that several foreign countries have adopted television standards based upon those of the United States with only minor variations.

In 1941 the same group of engineers who recommended standards for black-and-white television considered color television, and reported that the state of the art was such that it was undesirable at that time to attempt to set color standards. Five proposed color systems, intended to operate in a 6-mc channel, were analyzed by the NTSC. Four of these were variations of the CBS system with the principal differences in lines, frames, and type of interlace.

Much has been made of the fact that members of four of the NTSC committees were overwhelmingly in favor of color television as demonstrated by the Columbia Broadcasting System. Overlooked, or at least underemphasized, are some other opinions expressed by these engineers at the same time. By substantial majorities these engineers voted against adoption of color standards at that time and expressed the belief that black-and-white standards should not be influenced by color television considerations.

At the risk of digressing, I would like to comment that the question of whether any one prefers color in television is as elementary as "do you like a sunny day?" Of course people like color, and engineers, I believe, are usually regarded as people. But when it comes to a question as to whether color television is feasible, economical, and workable, there is no one better qualified to voice an opinion than the engineer who can consider color television in relation to all of the complicated and technical aspects of the transmission and reception of information by means of electronic equipment.

Following the war, in 1946 the first formal request that color television standards be established was made to the Federal Communications Commission. At this time the Radio Technical Planning Board—another and newer committee of industry engineers—stated that "adequate standards for color television for a 6-mc channel cannot be established at this time." The FCC apparently concurred in this opinion and denied the request that color television standards be adopted.

Again in 1948 the Commission reopened hearings on color television, and another industry committee—the Joint Technical Advisory Committee, sponsored by RMA and IRE—stated that in its opinion, based on evidence submitted by various subcommittees, it was "impracticable to set up commercial standards for color television in the present state of the art."

The most recent FCC hearing on color television in 1948–1949, I am sure, is fresh in your mind and the events need not be recounted. It should be noted, however, that a majority of the industry's engineers were in agreement that a compatible, acceptable color system of television could be developed within a reasonable length of time. This testimony was not accepted by the Commission, as it adopted the CBS field-sequential system instead. Now, of course, actual color broadcasting on the CBS system has been delayed by court action, and the present industrial mobilization program with its attendant shortage of critical materials has, at least for the time being, made color an academic question.

Thus, as I have pointed out, the engineers have been in the center of the color-television controversy for the past ten years, and until recently their opinions have been respected and their judgment followed. The great contribution television is making and will continue to make in the fields of entertainment, education, and commerce is sufficient evidence of the soundness of this engineering judgment. When practical color television is ready for the public, the industry's engineers will again be the key men in bringing this additional service to the public at the lowest cost and with a minimum of dislocation.

If wholesale indictment of your great profession, which I quoted a few minutes ago, were not so damaging and unfair, it would be ridiculous. You have given to the world so much in the field of electronics that I am confident the public does not concur in the views expressed by this government official. Your work, individually and collectively, and that of your great Institute, utterly refutes these unjust charges. I only urge you to continue to be guided in the future, as in the past, by your own high standards, and I venture to offer you this credo and code for engineers:

Differences of engineering opinion can and should honestly exist. Unanimity of engineering judgment is not necessarily desirable.

Changes in engineering opinion resulting from the passage of time and the gaining of experience are healthy, and normally to be expected.

It is improper and futile to ask engineers to reach final and valid conclusions at too early a stage in the development of new systems or equipment.

It is not only ethically correct but essential in the public interest that engineers shall duly weigh the economic factors involved in their planning work. Engineers are charged not only with the production of devices, but also with the selection of the least expensive devices which, under existing and likely future circumstances, will adequately serve the public at the lowest cost. In other words, engineering includes common sense.

There is conclusive evidence that engineers as a group and individually have at least as high moral, ethical, professional, and mental standards and qualifications as any other useful group of citizens.

There is also unchallengeable evidence that engineering societies are openminded and fair in their solicitation and selection of papers for publication in their journals, in their committee membership from various companies, and in their honorary awards.

We in the RTMA are rightly proud of what our industry has done for America in terms of communications, radio broadcasting, television broadcasting, military security, and host of other services. We know that the national welfare and prosperity have been furthered by our efforts and achievements.

You in The Institute of Radio Engineers have labored mightily and successfully to make your aims and ours come true. You properly share our pride of accomplishment in the public interest.

May both our organizations long continue to co-operate effectively for the happiness, comfort, and safety of our nation.

Before closing, I would like very briefly to sum up how it appears to me the present mobilization program will affect the radio-television industry. Our latest information indicates that military electronics production will reach a peak rate of $2.5 billion in the fall of 1952, and thereafter will decline to an annual rate of about $1.5 billion. In this connection, it should be borne in mind that military production dollars have about half the impact on our industry as civilian production dollars. This is for a variety of reasons, but particularly because a considerable portion of special and elaborate mechanical gear is obtained from manufacturers not generally considered a part of our industry.

These figures on military electronics production indicate that our industry will not be so heavily loaded with military contracts but rather it will be able to maintain a substantial amount of civilian production, even at the peak of the military output—except in the unfortunate event of an all-out war. Apparently, only the shortages of certain critical materials will prevent manufacturers from turning out as many radio and television sets as their plant facilities and military orders will permit. It is, therefore, highly important that government officials make provision in their planning for the healthy continuation of our civilian economy, for we do not know when all of the present manpower and production facilities in our industry may be needed for the nation's defense.

# A Dual-Diversity Frequency-Shift Receiver*

D. G. LINDSAY†

> Continuing the planned program of publication of valuable tutorial material, there is here presented a paper of this character. It has appeared in the *Proceedings of the Institution of Radio Engineers, Australia*, and is available to these PROCEEDINGS OF THE I.R.E. because of the close and co operative arrangements existing between these two Societies. It has also received the approval of the Subcommittee on Tutorial Papers of the IRE Educational Committee.—*The Editor.*

*Summary*—**Major improvements have been made in radio-telegraph transmission over a period of years, but the improvement resulting from signaling by frequency shift instead of interruption of the carrier wave has been achieved in practice only recently. These matters are discussed, and associated with a number of references to other work in this field. A description is given of dual-diversity receiving equipment, employing audio-frequency filters for separation of "mark" and "space" signals. The equipment is intended for use on direct-printing, commercial telegraph circuits.**

## I. INTRODUCTION

IN THE EARLY days of radio telegraphy, the method of operation involved aural reception and hand-key transmission by skilled operators at average working speeds seldom in excess of 20 words per minute, even under the best transmission conditions. Because of the discriminating ability of the human ear, experienced operators could copy very weak signals, marred by noise or interference, with remarkable accuracy. The radio circuit systems available adequately fulfilled the service necessary in those days, but requirements soon arose for greater efficiency of use in many radio circuits: for instance, higher speed or more direct methods of conveying the message such as by direct-printing methods became desirable. To this end, the use of methods and apparatus borrowed from wire telegraph technique quickly revealed that the speeding up caused by the removal of the human limitation in Morse speeds, removed also one of the most valuable of the operator's contributions, that of discrimination between signal and noise. To be of much practical use for commercial purposes, a radio-telegraph circuit employing any of the "mechanical" substitutes for operators was found to require, by comparison, a much better signal-to-noise ratio. These requirements involved the development of improved apparatus, the use of high transmitting power, better aerial systems, and so on, resulting in increased costs for equipment and maintenance.

A definite limitation to the increase of signaling speeds on the low frequencies then in general use, resulted from the increased bandwidth required for proper transmission of the signaling characters and the difficulty of obtaining this, due mainly to the low decrement of the transmitting aerial systems. At about this time the possible implications of scientific theories and some practical investigation, into the propagation of radio waves over the earth's surface, gave rise to ideas that frequencies above 1,500 kc, previously considered useless, might in fact prove valuable, especially for long-distance communication. It was realized, at once, that the use of very much higher frequencies would almost eliminate the restriction on signaling speeds and development of these new ideas into practical apparatus and communication circuits proceeded at a remarkable rate. The provision of direct radio circuits between England and Australia was among the first of the major communication projects to employ the high-frequency range, and due to the foresight and skill of the engineers concerned proved to be remarkably successful from its inception.

The new communication medium was found to have many vagaries in transmission conditions and it seemed almost impossible to predict the reliability of any given communication circuit, although the practical results were generally very good, showing great improvement over the previous technique in economy and speed. Further careful investigation and analysis of practical results revealed many methods for further improvement, such as directive aerial systems (in use on the England-Australia circuit from its inception), diversity reception, transmission employing frequency diversity by small variations of the transmitted frequency, general improvement in apparatus, and the application of much information on the correct choice of operating frequencies. These developments were made, and applied to commercial systems progressively, over a period of years.

By this time the use of machine telegraph methods had gained wide acceptance for commercial circuits, the improvements in technique and apparatus providing signals sufficiently good in quality for this purpose, despite the further vagaries of the transmission medium as compared with the earlier circuits on low frequencies, while the search for further improvements went on continually. The keying methods in use up to this time were, almost universally, the transmission of the carrier wave for the "mark" code. It had been realized for a long time that transmission of the required intelligence by varying the frequency of the emitted wave would be possible, but it was not appreciated that some advan-

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tage may accrue by so doing. In this way the method of telegraph signaling by shifting of the carrier-wave frequency, instead of interrupting it, now gaining wide acceptance, may be compared with its near relation in the telephony field, frequency modulation, the advantage of which has taken a long time to be appreciated.

In the very early days of radio telegraph transmission, signaling by means of frequency shift was employed because of difficulties in interrupting the carrier wave. This was the case in Poulsen arc transmitters and the keying method involved the transmission of the normal frequency for the "mark" condition and a shift of frequency, by altering the tuned circuit values with a relay or similar device, for the "space" condition. In this case, the "space" signal was not used to convey intelligence and only created undesirable interference, although it has been stated that some operators were sufficiently skilled to be able to copy the "space" signal, when the "mark" or normal frequency was subject to bad interference, often by the "space" signal of some other transmitter.

The basic difference between the modern methods of employing frequency-shift transmission and this very early example is that the frequencies corresponding to "mark" and "space" in the former are both used to convey intelligence.

The first major use of frequency-shift radio-telegraph appears to have been made during World War II (Sprague, 1944, Buff, 1946, and Vanderlippe, 1947). The lack of trained code operators and the desire for expediting communications made the advantages of the frequency-shift system attractive and, generally, good results were achieved. The large improvement obtained, however, tended to force the application unduly and resulted in apparatus originally designed for on-off keying being modified by the addition of various units, often with unsatisfactory results requiring frequent adjustments. This was because of the unsuitability of the original equipment, mainly with respect to frequency stability. Endeavors to overcome this by various expedients in the frequency-shift apparatus itself did not prove entirely successful.

## II. GENERAL

### *Principle of Frequency-Shift Keying*

As used in the frequency range between 2 and 30 mc, frequency-shift keying involves the transmission of two slightly different frequencies for the "mark" and "space" signaling conditions. The amount of frequency shift employed may vary with desired applications; one value that has been frequently used is a total shift of 850 cycles, that is, plus and minus 425 cycles from the assigned frequency, the "mark" signal, by the generally accepted convention, being the higher frequency of the two. In comparing the method with on-off keying it may be said in general terms that the two methods are special examples of the two fundamental modulation methods, amplitude modulation and frequency modulation. On-off keying is equivalent to 100 per cent amplitude modulation of a carrier with square waves and, similarly, frequency-shift keying of the type described is equivalent to frequency modulation of a carrier with square waves.

### *General Methods of Transmission and Reception*

As might be inferred from the details given above, frequency shift transmitters and receivers require a high degree of stability in the frequency determining elements, both for the amount of frequency shift and the nominal value of the carrier frequency. In practice, depending on the particular arrangement adopted for the apparatus, the value of frequency shift may be allowed to change a few hundred cycles due to variations of the shift itself, or the carrier frequency, or both, unless automatic-frequency control or bias-correcting circuits are used, in which case variations in carrier frequency of a relatively slow nature, up to perhaps 2,000 cps, may be tolerated. Where it is desired to keep the channel width small total variations of ±200 cps or less are desirable. These requirements are reflected in the design of both transmitting and receiving equipment, as the achieving of such stability is by no means an easy matter, although it can be obtained by the application of existing technique.

Of several methods that have been developed for use in the frequency-determining portions of a frequency-shift transmitter, one particular method has been widely used. This method involves the direct-frequency modulation of a low-frequency oscillator on a frequency of about 200 kc by the signaling code, the mixing of this output with a crystal oscillator and selection of the upper sideband frequency of the combination, followed by frequency multiplication and amplification to reach the desired final frequency and power output. Apart from the keying and frequency-control sections, the transmitter is generally quite conventional and equipment previously used for on-off keying is frequently employed, sometimes with a small reduction in power output because of the continuous duty-cycle. Both low-frequency oscillator and crystal oscillator are ordinarily temperature-controlled and a stability of a few cycles per megacycle can be maintained over periods of many hours with such equipment, accompanied by small variations of the amount of frequency shift in the order of a few cycles. Descriptions of frequency-shift transmitter exciter units using this general method are given in several of the references listed at the end of this paper.

Numerous methods have been developed for the reception of frequency-shift signals and it is safe to say that many more will be developed; differences in some cases are of a detailed nature but nevertheless may be important in obtaining the best possible performance in practice. The technique can be compared in some respects, again, to the developments in frequency-modulation receivers for telephony, where the variety of methods is ever increasing. In frequency-shift reception there is as yet no ideal system which makes all others obsolete; rather there are a number of methods, some differing only in detail., which have their own advantages and disadvantages, the choice of any one of these depending largely on the desired application. The choice of systems, for instance, may well be different for high-speed telegraph circuits using tape recorders, from a circuit in which it is desired to use direct-printing teletypewriters where the signaling speed is much lower.

The actual receiver units employed for frequency-shift reception are usually fairly conventional and not greatly different from those employed in on-off telegraph reception. Good frequency stability is a basic requirement; other design differences may result from the differing requirements of operating frequencies and similar factors. It is usual to employ two receivers in space diversity, that is, operating from aerials spaced apart five wavelengths or more, employing common oscillators for frequency determination, a common automatic-gain-control system and some means of combining the final outputs in the succeeding detector apparatus. It has been found that an improvement of the order of 10 db is obtained from the use of two receivers in diversity where noise is a limiting factor, and greater gain where multipath fading is a limitation (Petersen, Atwood, Goldstine, Hansell, and Shock, 1946). Also, there is only a small amount of further improvement by using three receivers in diversity; in this respect, frequency-shift systems are different from on-off systems, as in the latter a considerable improvement is effected by using three receivers in diversity. A number of points of interest in connection with the design of receiver units are covered in a later section of this paper where a complete equipment which has been developed is described.

It is in the means employed for dealing with the frequency-shift signals after passage through the receiver units that the main differences in the various methods appear. It is convenient, for the objectives of this paper, briefly to describe and contrast two methods which have found considerable application. These are referred to as the "audio filter" and "discriminator" methods.

In the former method, the receiver units are so designed that audio-frequency outputs are obtained from the receivers at a suitable frequency, spaced, of course, by the amount of frequency shift. If we take as an example a shift of 850 cycles, the two frequencies could be 2,125 cycles ("mark") and 2,975 cycles ("space"). These frequencies are applied to a filter for noise reduction, to a limiting amplifier to eliminate level variations, and are then applied to separating filters which comprise narrow band-pass filters centered around the "mark" and "space" frequencies, respectively. Further amplifier and rectifier stages (one each for "mark" and "space") then deal with the separated signals and derive the desired output for operation of the telegraph machine.

In the "discriminator" method the signal, varying in frequency by the "mark-space" difference, is usually taken from the receiver at the intermediate frequency and is then amplified and converted to a lower frequency, in the range 30 to 50 kc, further amplified and limited, and finally applied to a linear discriminator of the type often used for frequency-modulation telephony. The output of the discriminator is followed by a low-pass filter, chosen to suit the keying speed, and comprises nearly square waves of opposite polarity for mark and space, which are arranged to derive the output signals for operation of the telegraph machine.

Although the method in which the audio filters are used is commonly referred to as the "audio-filter" method, it must be appreciated that these audio filters constitute a frequency discriminator and the main difference between this and the so-called "discriminator" method is that the former is a *nonlinear discrminator* while the latter is a *linear discriminator*.

An advantage of the "audio-filter" method is that it can be so designed that moderate amounts of frequency drift do not cause telegraph bias. Provided the mark and space frequencies fall within the pass bands of their respective filters, it does not matter whether they are exactly centered or not. Further, during a period in which the communication circuit is idle and the transmitter is radiating a continuous mark condition, as is the normal convention, the output of the mark circuit is maintained and will hold the recording apparatus in a nonoperating condition. One of its disadvantages lies in the fact that it must be designed for a definite value of frequency shift. There is some latitude in the shift which can be handled, but it will be realized that the centers of the mark and space filters will normally be arranged to be separated by the expected shift. If designed for a shift of 850 cps the filters would not be optimum for a shift of, say, 500 cps. Most transmitter arrangements employed for frequency-shift circuits can be adjusted readily for various values of frequency shift but the inflexibility of the receiver in this respect is nevertheless a disadvantage. A second disadvantage is that the filter cutoff characteristics place a limit on the keying speed due to "ringing" effects. This method was originally designed to work with a fixed shift of 850 cps and for a definite telegraph speed in the vicinity of 50 bauds, corresponding to a speed of about 60 words per minute, the usual maximum working speed on teletypewriter circuits employing the special 5-unit code for these systems. For this particular service there is no doubt that the method gives very satisfactory results.

The linear discriminator method was initially developed to accommodate transmitter and receiver drift which could not be handled by the nonlinear discriminator system. A linear discriminator can be designed so that it has a long linear characteristic sufficient to accommodate the frequency shift plus the expected drift, assuming the receiver portion has an appropriate selectivity characteristic. The difficulty with this method is that when drift does occur, some standing direct current appears in the discriminator load network, directly proportional in amount to the drift from the true center frequency of the network, thus causing a bias to either "mark" or "space." Various circuit arrangements adopted to eliminate this bias tend to introduce other difficulties, one of which is a risk of false operation due to noise during idle periods, when the transmitter is signaling a continuous "mark." There is some evidence that the linear discriminator system is actually less tolerant of frequency variations than the "audio filter" system, even where bias correction is applied. Automatic-frequency-control arrangements are not usually applied in this system and are, in fact, somewhat difficult to apply, whereas such systems are frequently used, and can be more easily arranged, with the "audio-filter" system. The major advantages of the linear discriminator method are first, that a wide range of frequency-shift values can be accommodated, and second, there is no particular limitation on the telegraph speed.

It is emphasized that the details given above are a considerably over-simplified picture of the two methods; it is not completely fair to the merits of either, as detail variations that have been used in each, particularly in diversity-combining circuits, make it difficult to draw a sharp line of demarcation between them. There is, as yet, insufficient evidence of either a theoretical or practical nature to say definitely that any one system is the better; each has its advantages and disadvantages, as stated earlier in this section. Moreover, it is not the purpose of this paper to discuss these matters in full detail. Some of the references given at the end of the paper treat this portion of the design problem in greater detail, although there is some divergence of opinion, indicating that only further development of more extensive practical tests will clarify the position. Considerations such as ease of operation, ability to change readily existing on-off keying equipment and similar items, can also influence the choice of a particular method.

## III. IMPROVEMENT FROM THE USE OF FREQUENCY-SHIFT SYSTEM

### *Considerations of Signal, Noise, and Fading*

The frequency-shift system does not show any advantage over on-off keying for operation over radio-telegraph circuits that provide stable conditions and adequate signal levels similar to those on wire-telegraph circuits. In the case of long-distance radio-telegraphy in the high-frequency range, however, frequency shift shows a marked advantage. It is not possible to express quantitatively, by a single figure-of-merit, just how much better is the frequency-shift system in practice. This is because of the rapid fading and high noise conditions, multiple-path fading effects and so on, which commonly prevail in the high-frequency range. The results of theoretical treatment, laboratory tests, and field trials yield a unanimous verdict that there is a major improvement, but it is only possible to express this quantitatively under certain assumed conditions.

The result of extensive field trials are reported in a paper by Petersen, Atwood, Goldstine, Hansell, and Schock (1946). These tests were conducted over a considerable period, employing error counts to determine quality of a particular circuit, and adjustment or extrapolation of transmitter power to equality of error count was employed to give a comparison in terms of transmitted power, or field strength. These results are, of course, an average, not representing any one specific condition. It was found that a dual-diversity frequency-shift system, as compared with a three-receiver diversity system using on-off keying, showed an equivalent power gain of 11 db. The results also indicated that the gains caused by diversity reception (as compared with a single channel), amounted to approximately 10 db for both frequency-shift and on-off keying (two and three receivers, respectively) when noise is the limitation, with considerable diversity gain in the frequency-shift case where multipath distortion and not noise is a limitation. The gain caused by diversity reception where the limitation is multipath distortion cannot be expressed in terms of a power gain, as errors independent of power are caused.

In a paper by Ruddlesden, Forster, and Jelonek (1947), an interesting theoretical treatment is given of the improvement in signal-to-noise ratio by using the frequency-shift system. With the help of a number of simplifying assumptions, values of 10 and 14 db are obtained for the improvement, using audio-filter and discriminator-conversion methods, respectively, for a single channel. If it is inferred that double-diversity reception eliminates most of the effects caused by fading, the agreement between these figures and those quoted above for field trials is of interest.

A very detailed paper by Davey and Matte (1948) gives figures on the improvement caused by the frequency-shift system where noise is the limiting factor, of the order of 3 to 6 db, again for a single-channel system. These results do not agree closely with those quoted above. Greater improvements are indicated, however, for smaller amounts of signal distortion.

It is in dealing with rapidly fading signals, such as are usually encountered in hf telegraph circuits, that the circuits and the frequency-shift system shows to its greatest advantage. In considering this it is necessary for purposes of description to distinguish between non-selective or "flat" fading and selective fading, sometimes referred to as "multipath propagation." Selective fading implies propagation by such paths that two or more signal elements can arrive separated in time by amounts up to two milliseconds or more. This is significant since phasing conditions for slightly different frequencies can cause wide variations in level. Non-selective fading means that large level variations may occur without appreciable differences in path length so that frequencies a few hundred cycles apart do not fade by greatly different amounts. In practice, of course, both types often occur in combination over periods applying to communication.

Assuming, for the moment, that diversity reception is not being employed, considering nonselective fading and the on-off keying system, the effect of level variations, slow compared with the signaling speed, can be reduced by suitable automatic-gain-control (agc) circuits. In the case of rapid variations, the automatic-gain-control circuits can do little to assist and as the on-off system is relatively critical to level changes in the detected output, unless some compensating arrangement is adopted, failure will occur in hf circuits where the level variations are often extreme and rapid. Moreover, the failure will occur regardless of transmitted power. In a frequency-shift system, the agc circuits can have much smaller time constants, because of the presence at all times of either "mark" or "space" signals; efficient limiting amplifiers can be used and the system as a whole is not particularly critical to level changes, the frequency change being the significant factor. Some telegraph bias is produced if extreme level changes occur at the moment of frequency change between "marking" and "spacing" conditions, but in general the frequency-shift system is much more tolerant of nonselective fading.

With selective fading, again assuming that diversity reception is not used, the effect of large differences in path length in the normal on-off system is usualy a filling in of the spacing intervals with resulting telegraph bias; in many cases the distortion of characters is very severe. In frequency-shift systems, large time differences in the propagation paths (2 to 5 milliseconds) give rise to disturbances at the beginning of each signal element, caused by simultaneous arrival of "mark" and "space" signals resulting in a beat tone of frequency equal to the amount of frequency shift employed. As the frequency shift is generally several times greater than the highest, significant, modulating frequency which results from the keying signals, the action of the normal types of frequency-shift, detection arrangements largely reduces these effects. There is some divergence of opinion concerning the merits of frequency shift in relation to selective fading, but a recent paper by Bray, Lillicrap, and Owen (1947) gives a great deal of valuable data on this point, these workers employing a "fading machine" to perform tests under controlled laboratory conditions. Their evidence indicates that distortion of the signal elements in the usual frequency-shift systems can be much smaller than with on-off keying and the results are supported by field trials to which reference has previously been made (Petersen and others, 1946).

Diversity reception is commonly employed in hf radio-telegraph circuits, usually of the space-diversity type, as frequency diversity is wasteful in channel width. Diversity methods can improve greatly the performance of both on-off keying and frequency-shift keying systems, provided the diversity combining or selective method has the proper action and speed of response. This is especially important for frequency-shift keying (Lyons, 1946). With on-off keying the diversity action also helps to average rapid level changes to which this system is sensitive; for this reason, an appreciable improvement results from increasing the number of receivers providing the diversity action, from two to three. Since a frequency-shift system is tolerant of level changes, it could be expected that a smaller improvement would be obtained in changing from double to triple diversity and this is borne out in practice.

### *Bandwidth*

It is generally agreed that for a given signaling speed there is no great difference in receiving bandwidth requirements for on-off or frequency-shift keying. The "audio filter" method of frequency-shift reception does not appear to require a wider band than the "discriminator method" (Jones and Pfleger, 1946), but the difference is not of great practical significance as the failure point caused by noise does not seem to be affected.

For transmitters, the channel width occupied by a communication circuit operating at a given telegraph speed should be as narrow as possible; this matter is becoming of increasing importance in an already overcrowded spectrum and the interference created by any circuit must be reduced to a minimum. Contradictory though it may appear at first sight, the frequency-shift system is more economical in channel width and creates less interference than conventional on-off systems operating at equivalent speeds. With either system, keying by signal elements which are actually square waves would result in the occupation of a wide band and the creation of much undesirable and unnecessary interference. It has been found that considerable departure from the theoretical square-wave signal elements can be tolerated, provided the circuits pass harmonics up to the third of the effective modulating frequency (proportional to keying speed). Shaping of the signal elements by filter networks is employed for this purpose. In conventional on-off transmitters, however, it is usual to carry out the keying in low-level stages, allowing the successive higher power stages to cut off as the driving energy is removed. These later stages are ordinarily operated at the highest possible efficiency, that is under class-C conditions. In these circumstances well-shaped keying signals can become considerably distorted in their passage through such amplifier stages, giving rise to undesirable steep-sided transients which spread the band over which interference is created. The obvious alternatives of carrying out the keying and shaping of signal elements at a high power point, or employing class-B linear stages after a low-power stage which is keyed, are not at all attractive from the practical viewpoint. It is reasonable to state that insufficient attention has been given in the past to these factors and that valuable channels are not being used to the best advantage.

By contrast, the frequency-shift system enables the proper shape of keying signals to be preserved in passage through successive stages, since there is always a signal present and the small variations in frequency from "mark" to "space" pass easily through circuits as normally used.

Thus, it appears reasonable to expect a considerable reduction in side-frequency amplitude with frequency-shift systems, as compared with conventional on-off systems. A paper by Wickizer (1947) gives both theoretical and practical data on this matter; reduction in side-frequency amplitudes measured at one kilocycle from the carrier frequency of at least 20 db, are indicated. Similar conclusions are reached in a paper by Davey and Matte (1948). These considerations and data disclose a further important advantage for frequency-shift keying.

In telegraph communication circuits of either type relatively narrow bands can and should be used. Frequency stability of both transmitter and receiver can have a large effect, therefore, on the effective bandwidth occupied. There is every reason to demand the best possible stability in either on-off or frequency-shift systems so that the band occupied shall be no wider than necessary. Frequency-shift systems are much less tolerant of small frequency changes than are on-off systems, but these more stringent requirements can be met by the use of well-known techniques. What might be considered a disadvantage of the frequency-shift system, therefore, is in fact in its favor because it forces the attainment of better stability and further emphasizes the advantages in respect to reduced spreading of the band by keying, as outlined above.

### *General Comments*

The above information is in the nature of a critical review of technical opinion, as expressed in available data, together with comments and explanation. While there is some divergence of opinion about the advantages offered by frequency-shift keying, and difficulty of expression in quantitative terms and in an unqualified manner, there seems little doubt that there are many advantages. There are competing alternative systems, such as the use of a number of telegraph channels, possibly combined with telephony channels, in a single-sideband system, but the frequency-shift system appears attractive for the requirements of many communication circuits, especially where the period of reliable working per 24 hours is to be as long as possible. There also seems no reason why time-division multiplex systems cannot be applied to obtain multichannel facilities (Bray, Lillicrap, and Owen, 1947, and Petersen, and others, 1946). There is considerable evidence to indicate that many radio-telegraph circuits in use overseas are being changed progressively to frequency-shift working and this process is likely to continue, assisted by development work which is also continuing in an effort to improve further existing apparatus and technique. Present apparatus for frequency-shift systems is considerably different from that used in on-off systems, particularly at the receiving end, but is not much more complicated or expensive, and may be simplified as the result of further development.

## IV. DESCRIPTION OF A FREQUENCY-SHIFT RECEIVER

### *Initial Requirements*

The equipment to be described is intended for use in radio-telegraph teleprinter services over varying distances employing frequencies in the range 1.5 to 30 mc. Teleprinters employ a five-unit start-stop code in which all signal elements occupy nearly equal time intervals, approximately 22 milliseconds, resulting in a relatively low signaling speed of about 50 bauds corresponding to 60 wpm. For this purpose the "audio filter" method of separating the "mark" and "space" signals is suitable, resulting in the necessity for choice of a nominal value of frequency shift, given as 850 cycles in this instance. Other requirements giving a basis for design were:

(a) Equipment to be self-contained and rack-mounted, preferably in a single rack not more than 7 feet 6 inches high.
(b) Operation from single-phase power mains of 200 to 260 volts, 40 to 60 cycles.
(c) Double-diversity receivers required with provision for connection to balanced aerials of 200 ohms impedance, or 75 ohms unbalanced.
(d) Equipment to operate with a minimum of attention once set up for a particular circuit. Arrangements for automatic-frequency control desirable.
(e) Receivers to provide for three pre-set frequencies in the range 1.5 to 30 mc, with crystal control of the frequency-changing oscillator; or alternatively, selection of a variable oscillator for use in rapidly setting up a new circuit for which a suitable crystal may not be available.
(f) Test equipment to be provided for checking frequency drift and for carrying out tests on the audio-frequency portions of the equipment.

### *General Arrangement*

As indicated in the previous section, it was desired to mount the equipment in a rack, using standard 19-inch panel-mounting units. This involves division of the equipment into a number of units each of which can be mounted at convenient points in the rack structure. The arrangement adopted is shown in block-schematic form in Fig. 1 and the disposition of the units in the cabinet-type rack in Fig. 2. The units indicated in these figures are separately described in later sections; this section deals only with their general arrangement and functions.

Referring to Fig. 1, the aerial coupling unit, two receivers and the main oscillator are seen on the left side of the figure. The aerial coupling unit provides terminations for the aerial feeders (balanced or coaxial) and coaxial connectors for output to the receivers. Where balanced aerials are used, switches enable the selection of coupling transformers, providing the 200- to 75-ohm matching and balance-to-unbalance connections required. Two coupling transformers for each receiver are fitted, covering the frequency ranges 1.5 to 6.7 mc and 6.7 to 30 mc.

The receivers are arranged for an unbalanced 75-ohm input circuit and when the aerial feeders are 75-ohm coaxial cables, the aerial coupling-unit switching arranges a direct connection to the receiver input circuits. As is common practice for diversity receivers, the frequency-changing oscillator is common to the two receivers and the unit containing this oscillator (termed the main oscillator unit), also includes the beat-frequency oscillator and its afc circuits, again common to the two receivers. Apart from this arrangement of oscillators, the receivers are, for the most part, of conventional superheterodyne type.

Output from the receivers at audio frequency (2,125 and 2,975 cycles for "mark" and "space," respectively) is taken to separate limiter-detector units for each channel. These units each contain an input filter, limiting amplifier, "mark" and "space" separating filters followed by a further amplifier stage, rectifier and relay valves, and a differential noise-suppressor circuit.

The output from these units, comprising dc pulses corresponding to "mark" and "space" signals, connects with the relay unit where the two channels combine at the coil of the keying relay. The keying relay is arranged to send dc to the output line derived from the rectifiers and a power transformer included in the relay unit.

The unit termed "frequency deviation indicator and vf keying unit" provides for indication on the center-zero meter of the deviation of "mark" and "space" frequencies from their nominal values of 2,125 and 2,975 cycles and a separate section for supplying test "mark" and "space" signals at these frequencies. It is employed for initial setting up of the equipment on a particular circuit and for observation during operation of any drift in center-frequency adjustments.

The other units shown in Fig. 1, comprising meter panel, jack strip, power switching panel, and terminal panel provide for test measurements of voltages, currents and audio-frequency levels, patching of unit inputs and outputs, control of mains supply circuits, and connections to external circuits, respectively.

Each unit which requires power-supply arrangements contains its own power transformer, rectifier valve, and filter circuits, and the mains input is controlled by separate switches on the power switching panel.

Fig. 2 shows the arrangement of the various units in the rack assembly, and also indicates the vertical space occupied by each. All major units are of recessed-chassis construction in which valves and large components project horizontally from the rear of the chassis, the open side of which faces the front, contains all minor components, and is covered by a readily removable panel. The rack is a metal cabinet, approximately 14 inches deep, with a hinged door over the full height at the rear. This method of construction of the individual units and assembly into the rack results in an equipment of good appearance and accessibility. The construction and components are such as to ensure reliable operation under extreme climatic conditions.

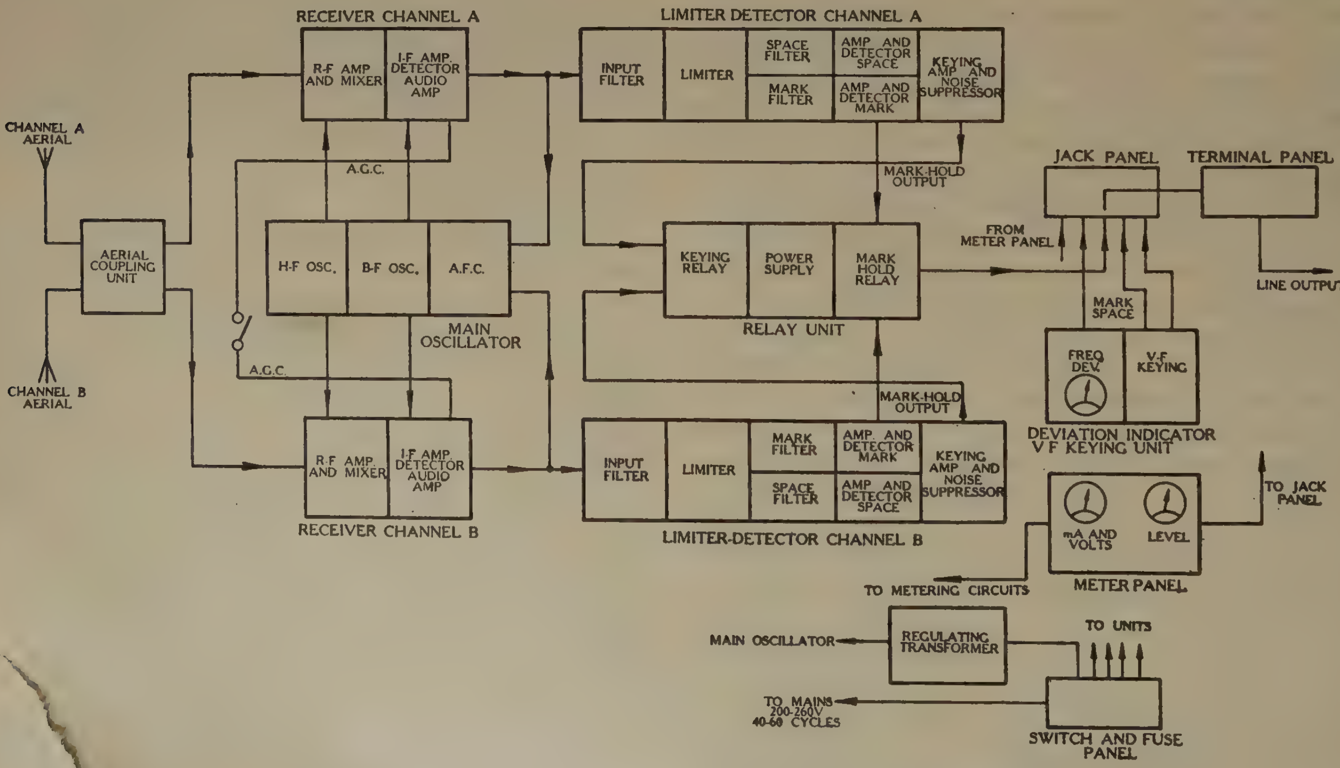
RECEIVER CHANNEL A
R-F AMP AND MIXER
I-F AMP. DETECTOR AUDIO AMP
LIMITER DETECTOR CHANNEL A
INPUT FILTER
LIMITER
SPACE FILTER
MARK FILTER
AMP. AND DETECTOR SPACE
AMP. AND DETECTOR MARK
KEYING AMP. AND NOISE SUPPRESSOR
CHANNEL A AERIAL
A.G.C.
MARK-HOLD OUTPUT
JACK PANEL
TERMINAL PANEL
AERIAL COUPLING UNIT
H-F OSC.
B-F OSC.
A.F.C.
MAIN OSCILLATOR
KEYING RELAY
POWER SUPPLY
MARK HOLD RELAY
RELAY UNIT
FROM METER PANEL
LINE OUTPUT
MARK SPACE
FREQ DEV.
V-F KEYING
DEVIATION INDICATOR V F KEYING UNIT
CHANNEL B AERIAL
A.G.C.
MARK-HOLD OUTPUT
R-F AMP. AND MIXER
I-F AMP. DETECTOR AUDIO AMP
RECEIVER CHANNEL B
INPUT FILTER
LIMITER
MARK FILTER
SPACE FILTER
AMP. AND DETECTOR MARK
AMP. AND DETECTOR SPACE
KEYING AMP. AND NOISE SUPPRESSOR
LIMITER-DETECTOR CHANNEL B
TO JACK PANEL
mA AND VOLTS
LEVEL
METER PANEL
TO METERING CIRCUITS
MAIN OSCILLATOR
REGULATING TRANSFORMER
TO UNITS
TO MAINS 200-260V 40-60 CYCLES
SWITCH AND FUSE PANEL

Fig. 1—Block-schematic diagram of frequency-shift receiving equipment.

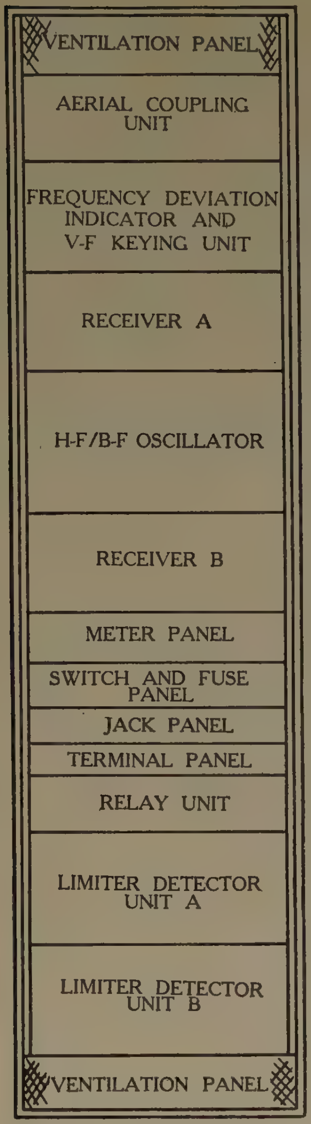
VENTILATION PANEL
AERIAL COUPLING UNIT
FREQUENCY DEVIATION INDICATOR AND V-F KEYING UNIT
RECEIVER A
H-F/B-F OSCILLATOR
RECEIVER B
METER PANEL
SWITCH AND FUSE PANEL
JACK PANEL
TERMINAL PANEL
RELAY UNIT
LIMITER DETECTOR UNIT A
LIMITER DETECTOR UNIT B
VENTILATION PANEL

Fig. 2—Dual-diversity receiving equipment rack layout.

### *Receivers*

Fig. 3 is a block-schematic diagram of the stage arrangement employed in each of the receiver units. Miniature valves are used throughout, except for the rectifier providing the high-voltage supply, because of their convenience and improved characteristics. Two amplifier stages at signal frequency are employed to achieve good signal-to-noise ratio and image response, and the full frequency coverage of 1.5 to 30 mc is divided into three ranges, normally 1.5 to 5 mc, 5 to 15 mc, and 15 to 30 mc, the desired range being selected by a switch on the front panel. There are no tuning controls of the conventional type, each coil circuit covers the range indicated with adjustments accessible from the rear of the unit. It is intended that the receivers will be present to the desired operating frequencies anywhere in the range indicated and the particular frequency required selected by the switch referred to above. If more than one of the desired operating frequencies falls in one of the ranges indicated, coil assemblies can be readily varied accordingly.

Two rf amplifier stages are followed by a mixer, in which the heterodyne frequency is taken from the main oscillator unit, giving an intermediate frequency of 455 kc, which is amplified in two stages employing a total of five double-tuned transformers. A switch is provided on the front panel for control of the selectivity given by the IF amplifier and it has three positions termed Wide, Normal, and Crystal. These positions give selectivity characteristics for 1 drop in response of approximately 3 db of ±3.5 kc, ±2 kc, and ±0.9 kc, respectively.

A separate amplifier circuit and rectifier is used to derive agc voltages, which are applied to all preceding stages. This is done for two reasons; first, because this

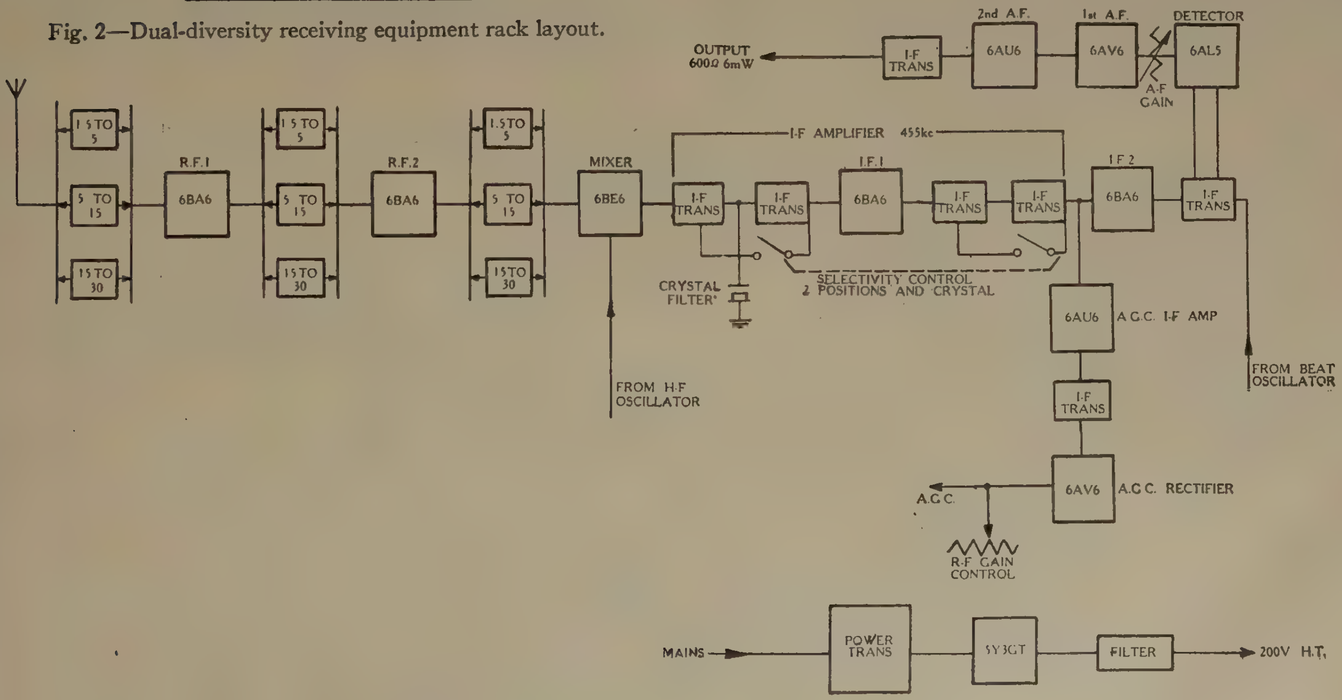
OUTPUT 600Ω 6mW
I-F TRANS
2nd A.F.
6AU6
1st A.F.
6AV6
DETECTOR
6AL5
A-F GAIN
1.5 TO 5
5 TO 15
15 TO 30
R.F.1
6BA6
R.F.2
6BA6
MIXER
6BE6
I-F AMPLIFIER 455kc
I-F TRANS
I.F.1
6BA6
I.F.2
6BA6
CRYSTAL FILTER
SELECTIVITY CONTROL 2 POSITIONS AND CRYSTAL
6AU6
A.G.C. I-F AMP
FROM BEAT OSCILLATOR
FROM H-F OSCILLATOR
6AV6
A.G.C. RECTIFIER
A.G.C.
R-F GAIN CONTROL
MAINS
POWER TRANS
5Y3GT
FILTER
200V H.T.

Fig. 3—Block-schematic diagram of the receiver.

method enables very good agc characteristics to be obtained, and second, because it ensures freedom from undesired effects on the agc caused by stray coupling from the beat oscillator. An rf gain control is provided, operating on the agc circuit in such a manner that the shape of the agc curve is not appreciably altered when the rf control is adjusted. An over-all reduction in sensitivity results only in a shift of the whole agc curve towards increasing input voltage for the beginning of action. Provision is also made for common connection of the agc circuits of the two receivers employed in the complete equipment. This is arranged through a switch enabling the common agc to be disconnected if desired.

A balanced detector circuit employing a 6AL5 double diode is used, enabling connection of the beat-frequency oscillator in such a manner that common coupling between the two receivers is minimized.

The two audio-frequency amplifier stages, employing 6AV6 and 6AU6 (triode-connected) valves, are quite conventional in arrangement and the output transformer is arranged to suit a 600-ohm load circuit. Provision is made for the connection of meters into the mixer circuit and both the agc and signal diode circuits, for assistance in initial alignment to a particular frequency.

### *High-Frequency and Beat-Frequency Oscillator*

Fig. 4 is a block-schematic diagram of this unit and it will be noted that there are three portions shown as independent of one another. The portion designated "power supply section" provides high-voltage and filament supply for the other two portions, termed "hf oscillator section" and "beat oscillator section," respectively. These latter two are actually independent sections electrically and are combined mechanically for convenience, as each operates in conjunction with the two receivers. The whole unit is referred to as the "main oscillator" in Fig. 1 and in the previous text.

The hf oscillator provides for selection of any one of three crystals, or of a variable tuned circuit covering the whole range in three steps. It is intended that the variable oscillator should be used only for initial alignment or in cases where a new circuit has to be set up quickly and no suitable crystal is available. Crystals used have individual temperature-controlled ovens, a complete crystal unit being approximately the size of a "G-T" valve. These crystals, the variable oscillator circuit, and the oscillator valves are contained in a further temperature-controlled oven, resulting in very close temperature control of the crystals (better than ±1°C), and control to about ±3°C of the variable oscillator circuit. Provision is also made for slight adjustments to the crystal frequencies (a range of about 0.01 per cent), by means of variable capacitors across each crystal, the adjustments being available on the front panel of the unit. The variable oscillator also has the main and vernier tuning controls on the front panel. Other circuits of the hf oscillator, outside the oven, comprise an untuned buffer stage and a tuned output stage, both employing 6AU6 valves.

The beat oscillator section provides for three conditions of operation: manual, in which the beat frequency can be manually adjusted; afc, in which it is controlled by afc circuits; and crystal, where it is maintained at 452–450 kc by a quartz crystal. The manual position is intended for use mainly in setting up a circuit, although it can be used to compensate for the inability of a crystal in the hf section to be adjusted precisely to the required frequency, as long as the variations are small enough to enable passing of the intermediate frequency thus created through the selective circuits of the IF amplifier. If the transmitter with which the receiver is working has adequate frequency stability, so that the variations in frequency caused by both transmitter and receiver over a working period are likely to be less than

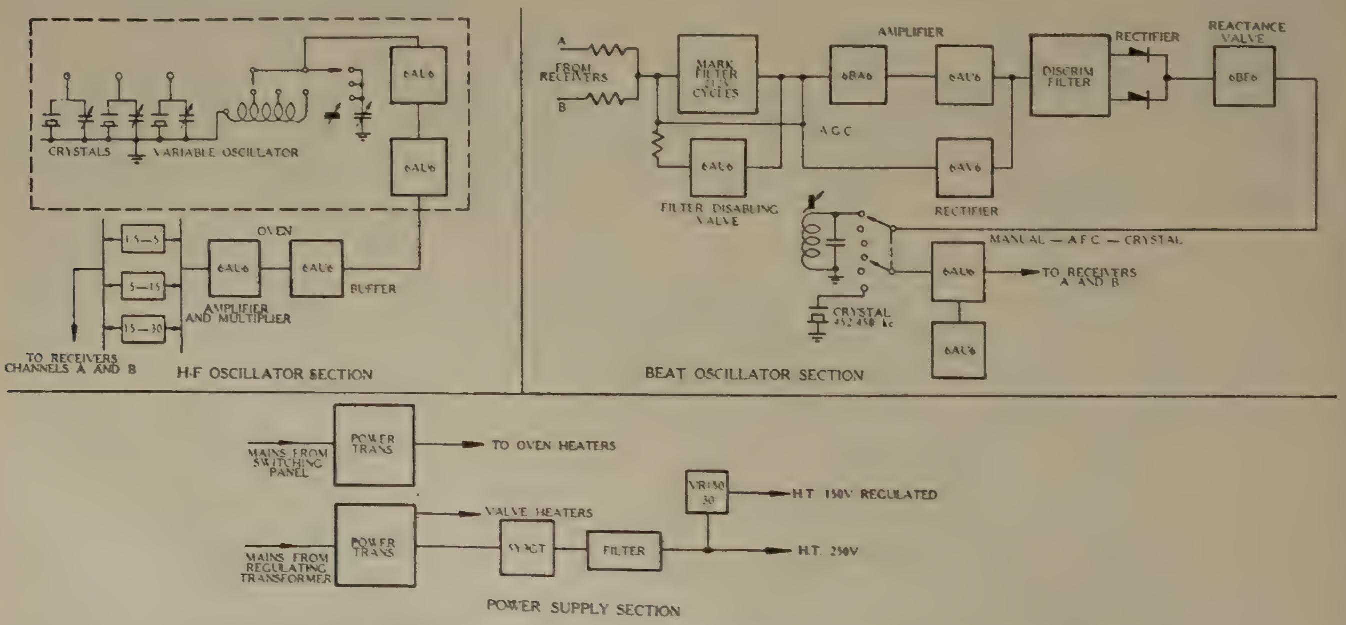
6AU6
6AU6
CRYSTALS
VARIABLE OSCILLATOR
OVEN
1.5—5
5—15
15—30
6AU6
6AU6
BUFFER
AMPLIFIER AND MULTIPLIER
TO RECEIVERS CHANNELS A AND B
H-F OSCILLATOR SECTION
A
B
FROM RECEIVERS
MARK FILTER CYCLES
AMPLIFIER
6BA6
6AU6
DISCRIM FILTER
RECTIFIER
REACTANCE VALVE
6BE6
AGC
6AU6
FILTER DISABLING VALVE
6AV6
RECTIFIER
MANUAL — AFC — CRYSTAL
6AU6
TO RECEIVERS A AND B
CRYSTAL 452-450 kc
6AU6
BEAT OSCILLATOR SECTION
MAINS FROM SWITCHING PANEL
POWER TRANS
TO OVEN HEATERS
VR150 30
HT 150V REGULATED
VALVE HEATERS
MAINS FROM REGULATING TRANSFORMER
POWER TRANS
5Y3GT
FILTER
H.T. 250V
POWER SUPPLY SECTION

Fig. 4—Block-schematic diagram of the high-frequency oscillator and the beat-frequency oscillator.

the variation allowed by the audio-frequency filters (that is, within about ±300 cycles), then the crystal position may be used. It is also useful during the initial setting up of a circuit. If the frequency variation caused by both transmitter and receiver is likely to be much greater than ±300 cycles and the rate-of-change of frequency is fairly small, then the afc position may be used. The afc circuits will compensate for drift up to about ±2,000 cycles if the circuit is set up correctly, so that an initial manual adjustment, referred to the crystal and frequency deviation indicator, gives the correct mark frequency. The afc circuit employs a mark filter for its input network (see next section and Figs. 6 and 7), deriving its input signal from both receivers through bridging pads. The circuit time constants are slow so that the correcting output voltage is maintained during keying of the transmitter, and the mark filter prevents any response from the space signaling condition. During periods in which no message is actually being transmitted, it is usual for the transmitter to radiate a continuous mark signal and, in these circumstances, the afc circuit continues to operate and correct for frequency drift. If, however, during a working period the transmitter ceases to radiate for some time and when it does recommence the frequency drift which has occurred is greater than 300 cycles, the mark filter would prevent the afc circuit from operating if it were not for the special by-pass circuit provided. The circuit uses a 6AU6 valve which is termed a "filter disabling valve" and is so arranged that a correcting signal is derived despite the wide variation in frequency, but once the frequency is again centered in the range of the filter, the valve is prevented from amplifying appreciably by a bias derived from a slow-acting, agc circuit. A range of ±2,000 cycles (approximately) is covered by this means. False operation of the afc circuit can be caused by interfering signals, or by inadvertent transmision of a space signal for a long period; in the first case, a complete failure of communication may occur; in the second, a recovery period comprising a few seconds of mark signal is required. Where the transmitter is of adequate stability and a crystal of proper frequency is available for the hf oscillator of this equipment, there would be no necessity for the use of the afc position. It can be very useful, however, if the transmitter is not sufficiently stable, or immediate operation on a new frequency is required, calling for the use of the variable hf oscillator.

The power supply section provides high-voltage and low-voltage power for operation of the two sections described above. A VR150/30 gas regulator valve is used to stabilize high potentials to the oscillators and the mains input is stabilized by a regulating transformer (mounted in the rack) to obtain the best possible stability from the variable hf oscillator which is suceptible to filament voltage changes. The heater supply to the temperature-controlled crystals and the main oven is provided from a separate transformer having an individual mains switch, so that the ovens may be left running continuously and other units switched off as desired. Pilot lamps are provided in the oven circuits to indicate operation of the thermostats.

Provision is made for metering circuits to provide for measurement of important voltage and current values in the unit, using the meter panel mounted on the rack.

### *Limiter-Detector Unit*

The general arrangement of the limiter-detector unit is shown in Fig. 5, applying to channel *A*. The unit for channel *B* is identical; the keyed dc outputs of the two channels combine at the relay unit and there is a common connection associated with the noise-suppressor circuit.

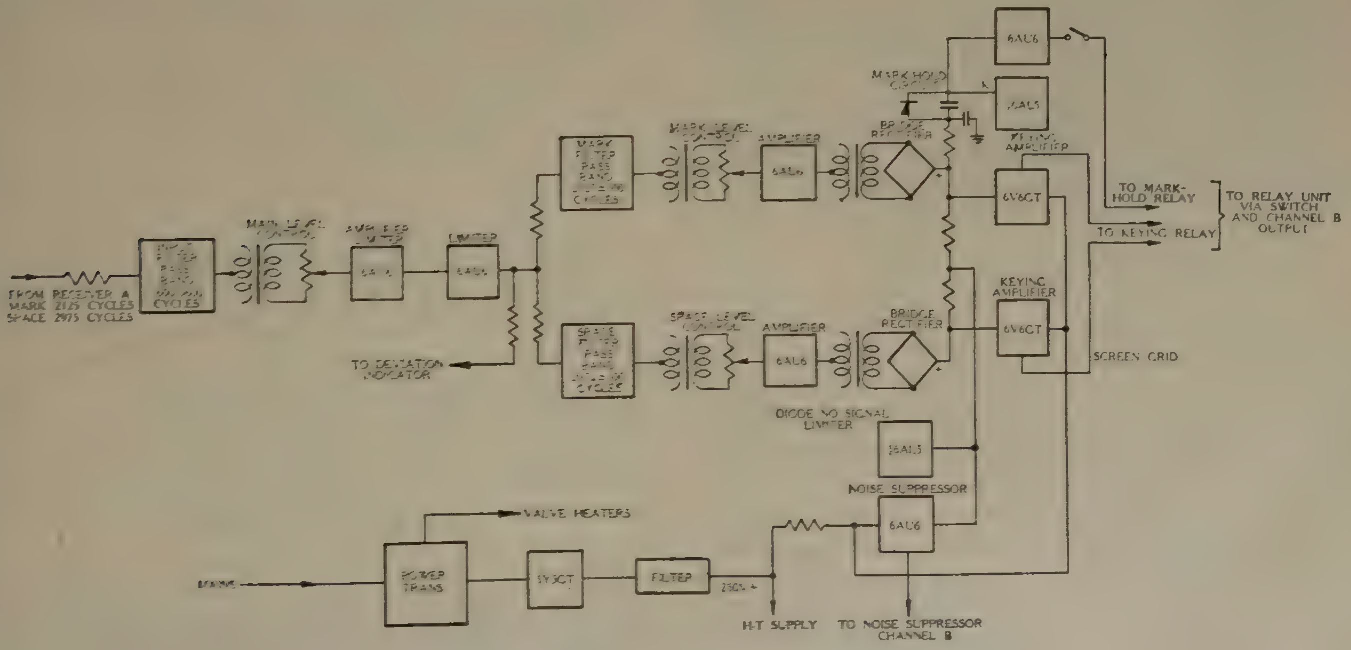
FROM RECEIVER A
MARK 2125 CYCLES
SPACE 2975 CYCLES
CYCLES
LIMITER
TO DEVIATION INDICATOR
MARK LEVEL CONTROL
AMPLIFIER
6AL6
BRIDGE RECTIFIER
MARK HOLD CIRCUIT
6AU6
½6AL5
KEYING AMPLIFIER
6V6GT
TO MARK-HOLD RELAY
TO KEYING RELAY
TO RELAY UNIT VIA SWITCH AND CHANNEL B OUTPUT
SPACE LEVEL CONTROL
AMPLIFIER
6AL6
BRIDGE RECTIFIER
KEYING AMPLIFIER
6V6GT
SCREEN GRID
DIODE NO SIGNAL LIMITER
½6AL5
NOISE SUPPRESSOR
6AU6
VALVE HEATERS
MAINS
POWER TRANS
5Y3GT
FILTER
250V +
H-T SUPPLY
TO NOISE SUPPRESSOR CHANNEL B

Fig. 5—Block-schematic diagram of the limiter-detector unit (channel A).

In the receiver units previously described there are not many features which are peculiar to the frequency-shift system; these receivers would be suitable for the on-off keying system if the initial requirements were similar. Conversely, receivers originally designed for on-off keying could be used with some modification, if their frequency stability were sufficiently good. The major differences in technique appear in the limiter-detector unit and its associated relay unit.

The limiter-detector unit deals with mark and space signals from the receiver output on nominal frequencies of 2,125 and 2,975 cycles, respectively. The difference between these frequencies is, of course, the deviation as transmitted, but the reason for the choice of 2,125 cycles for the mark frequency is of interest. In this unit the mark and space signals pass first through an input filter which is intended to allow for variations in the order of 300 cycles in both mark and space frequencies. As the nominal value of frequency shift is 850 cycles, this means that the input filter must have a pass-band of approximately: $2[(850/2)+300]$ cycles $=1{,}450$ cycles (or 725 cycles on each side of a nominal center frequency).

After passing through this filter, the mark and space signals are applied to a two-stage limiting amplifier to reduce level variations. Fig. 6 shows the input-output characteristics of the limiting amplifier. The output of the limiting amplifier is then applied to the mark and space filters which perform the separation of mark and space signals. These filters must each have a pass-band approximately 600 cycles wide (±300 cycles), but at the mean frequency (425 cycles above mark and 425 cycles below space), there must be considerable attenuation so that the separating process can be carried out properly. This mean frequency is chosen to be the third harmonic of 850 cycles, that is, 2,550 cycles, introducing considerable attenuation at this frequency. This fixes the center frequencies for mark and space at 2,125 and 2,975 cycles, respectively, and also indicates a suitable pass band for the input filter, 1,825 to 3,275 cycles. It will be recalled that multipath propagation can give rise to beats between the mark and space frequencies, that is 850 cycles, with production of harmonics of this frequency also likely. If the input filter is designed to cut off fairly rapidly below 1,825 cycles, it will obviously attenuate 850 cycles considerably and 1,700 cycles (second harmonic) to some extent; these facts, together with the noise reduction also attained, are the reasons for its use. The third harmonic falls at 2,500 cycles and is attenuated by mark and space filters; the fourth harmonic, 3,400 cycles, is appreciably attenuated by the space filter and the fifth is substantially eliminated. These are some of the considerations for the choice of audio-frequencies; others are, that any much higher frequency would introduce difficulties in aural checking and also require better percentage stability in the filters. These particular frequencies were chosen in some of the earlier equipment using an 850-cycle frequency-shift value and it is evident that a good choice was made.

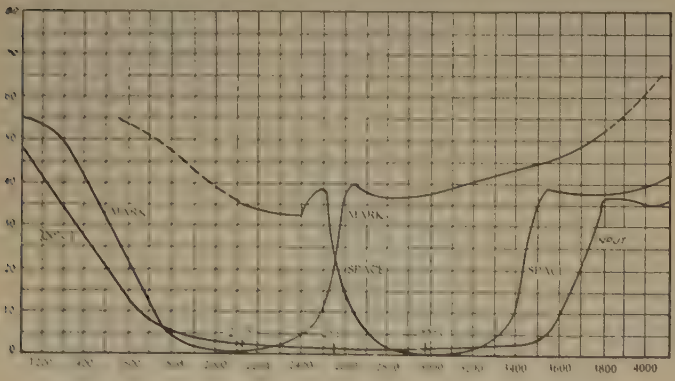

Fig. 6—Characteristics of filters.

The above nominal values for mark and space frequencies have been followed and result in practical filter characteristics as shown in Fig. 7. The circuit arrangement of each of the filters is shown in Fig. 8. It will be noted that the slope of the attenuation curve outside the nominal pass-band is not particularly steep, and also that a relatively large number of sections is em-

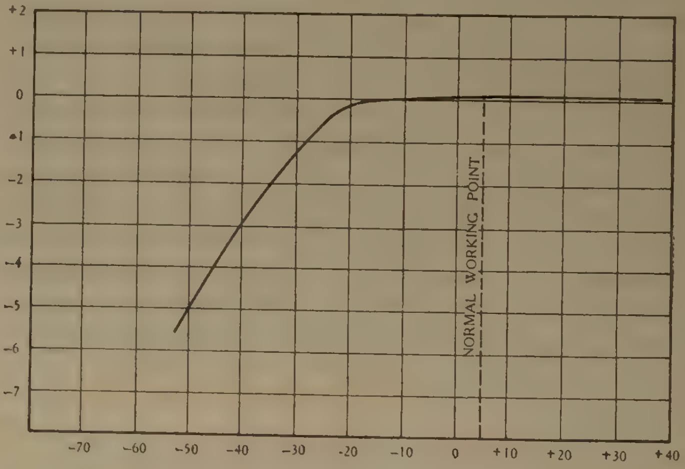

Fig. 7—Characteristics of limiter.

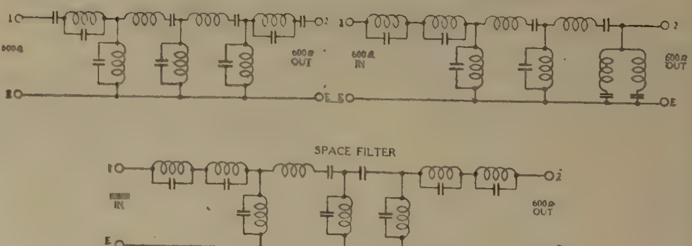

Fig. 8—Arrangement of filters.

ployed in each filter because the coils used are of relatively low-$Q$ value (25 to 30). This has been found advantageous as there is only a slight tendency to "ringing" in the filters at keying speeds double the normal working speed; correct resistance termination of the filters also assists in this respect. In addition, the filters have been designed to enable compensation for appreciable variations in the value of individual components to facilitate production in fairly large quantities. The price paid for these desirable factors is that the filters are somewhat large physically. Each type is contained in a metal tank approximately 5½ by 5½ inches (base), and 9 inches in height.

The mark and space filters are each followed by an amplifier stage, preceded by a level-adjusting control to enable adjustment of mark and space levels to equality. The amplifier stage for each circuit is succeeded by a transformer and a bridge-connected selenium rectifier, with the output polarity positive with respect to earth.

The outputs of the bridge rectifiers connect to the signal grids of two 6V6GT valves which act as a keying amplifier. The circuit is essentially that of a symmetrical dc amplifier with the valve cathodes connected together and to earth (ht negative) through a fairly large resistor. Application of the dc signal to one causes the other to cut off and drives an output current in one direction through half the relay coil; application of signal to the second results in a reverse action.

Considering the two channels, it will be obvious that, with normal signals in both, the keying amplifier of each channel will make approximately equal contribution to the output current operating the relay for either mark or space. If the signal fades in one channel it will be replaced by an approximately equal noise voltage, caused by the action of the limiting amplifier and receiver agc circuits. Note, however, that this noise voltage appears in *both* mark and space output circuits of the channel concerned. This results in an increase in voltage at the common connection of two resistors, the other ends of which connect to the signal grids of the keying amplifier valves (see Fig. 5). This change of potential, caused by fading of the desired signal and increase in noise, is employed to produce a suppressing action on the contribution of the affected channel to the output relay current. The potential change (positive for increasing noise) is applied to the signal grid of 6AU6 valve (noise suppressor) and results in a reduction of the screen-grid potential of the keying amplifier. In addition, a differential action is produced by connecting the cathodes of the noise-suppressor valves of channel $A$ and channel $B$ together and through a high resistance to ht negative. Thus, an increase in noise on one channel decreases its contribution to the output current and increases that of the other channel. When both channels are equally noisy, the output current decreases slightly but remains equal for both channels. Should one channel develop a fault prior to the keying amplifier, resulting in the elimination of both signal and noise, a diode circuit ("diode no-signal limiter") using half a miniature double-diode (6AL5) is employed to prevent the differential action of the noise suppressor from blocking the other channel also. The unaffected channel can then do its best to maintain the communication circuit. The complete noise-suppressor system is designed in such a manner that very rapid action is possible; a mark signal might be accepted from channel $A$ and the succeeding signal element from channel $B$ with $A$ making no contribution. The operating conditions of the keying amplifier valves are chosen to make the cut off, caused by reduction in screen potential, quite sharp and if one channel output is practically all noise the differential action results in a normal value of output current from the other channel (that is, its ordinary contribution is approximately doubled), and little or no output from the noisy channel. Assisted by the common agc circuits of the receivers this arrangement provides good diversity action and results in a marked improvement in communication reliability as compared with a single channel.

A further facility termed a "mark-hold circuit" is provided in each channel (see Fig. 5, near top of diagram). As indicated in an earlier section, it is the normal convention in frequency-shift systems for the transmitter to send a continuous mark signal when the circuit is "idle"; the first "space" is actually a "start" signal for the motor of the teletypewriter. However, when this convention is not followed, if the transmitter fails or is deliberately shut-down for a short period, it is possible that noise and interfering signals might cause false operation of the teletypewriter. To prevent this without requiring attention by either the receiver operator or the operator at the teletypewriter, a circuit is provided which allows normal operation but operates to signal a mark on the output line if normal keying, or the transmission of a continuous mark, ceases for a period in excess of about 0.25 second. The arrangement adopted involves the use of an integrating circuit associated with two rectifier diodes (one selenium rectifier and half a double vacuum-diode 6AL5—see above), and a 6AU6 valve, the anode current of which operates a relay. When the mark-hold relay drops out because of the failure of the signal, the relay contacts operate to send mark on the output loop, prevent the signal relay from chattering because of noise impulses, and operate a warning pilot lamp. The relay and pilot lamp are actually mounted in the relay unit and the anode currents of both mark-hold valves pass through the relay coil. A switch on each limiter-detector unit enables the mark-hold circuit to be made inoperative if desired; with very poor signals and high noise level, false operation is possible and extra attention from the receiver technician will be required to keep the circuit operating.

Provision is made for metering all important circuit voltages and currents using the meter panel mounted on the rack.

### *Relay Unit*

This unit, shown in Fig. 9, contains the keying relay, the mark-hold relay, power supply circuits to provide the desired line current, and a metering switch and meter providing for measurements of all important current values. Two further switches provide for the selection and combining of *A* and *B* outputs and the selection of the desired arrangement for keying to line.

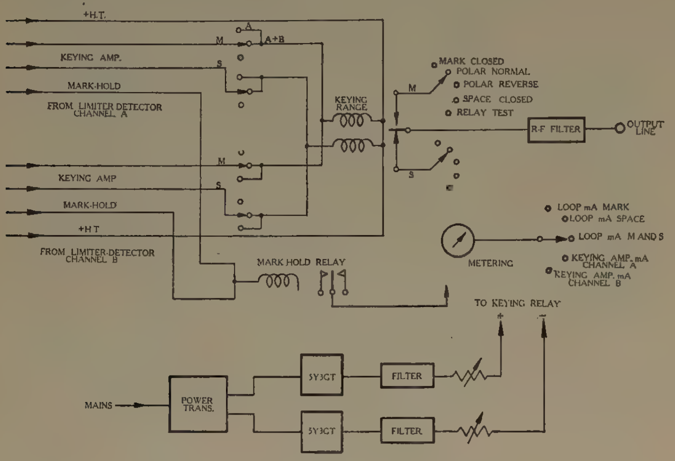
+H.T.
M
A
A+B
KEYING AMP.
S
MARK-HOLD
FROM LIMITER DETECTOR CHANNEL A
KEYING RANGE
M
MARK CLOSED
POLAR NORMAL
POLAR REVERSE
SPACE CLOSED
RELAY TEST
R-F FILTER
OUTPUT LINE
S
M
KEYING AMP
S
MARK-HOLD
+H T
FROM LIMITER-DETECTOR CHANNEL B
MARK HOLD RELAY
LOOP mA MARK
LOOP mA SPACE
LOOP mA M AND S
METERING
KEYING AMP. mA CHANNEL A
KEYING AMP. mA CHANNEL B
TO KEYING RELAY
+
−
MAINS
POWER TRANS.
5Y3GT
FILTER
5Y3GT
FILTER

Fig. 9—Block schematic, relay unit.

The output from *A* and *B* channel, limiter-detector units combines at the keying relay and it is desirable to have a means of using each channel separately for checking purposes; also, a defective channel may need to be isolated, since certain possible faults in one channel can react on the other. These facilities are provided by the switch shown on Fig. 9 to the left of the keying relay.

The keying relay is a conventional "side-stabel" high-speed type used with the coils connected series-aiding and the keying amplifier high-voltage supplies connected to the center-point of the winding. Normal signalling current is approximately 15 ma from the keying amplifier of each channel. The contacts provide for keying of the dc power supply to line via spark suppression filters and a switch which enables polar keying or on-off keying to be selected. In the on-off keying positions the local dc supply is not used, the relay contacts key the line and the dc supply would be provided at the remote end. In addition, the switch gives reversal positions (that is, reversed polarity for mark and space in the polar keying condition and current to line for space signals instead of mark in the on-off keying position). It is easily possible with the frequency-shift system to obtain reversed-keying conditions and although the point where the reversal occurs must be found and the final correction made there, it is convenient to be able to obtain a temporary correction in this manner. The normal current sent to line is approximately 60 ma for a 500-ohm loop circuit. This switch has one further position which places a potential of about 6 volts 50 cycles on the relay coil and terminates the line output in a resistor, removing all normal circuit connections. This enables checking and adjustment of the relay. Because of the importance of the current values in this unit a metering switch and meter are included. These provide facilities for measuring the keying-amplifier currents of each channel, or the combined value (in conjunction with the channel-selecting switch) and the loop currents keyed to the output line.

The mark-hold relay is a conventional telephone relay having a 10,000-ohm coil and contacts performing the functions described in the above section.

### *Frequency-Deviation Indicator and VF Keying Unit*

The arrangement of this unit is shown in Fig. 10. It provides for measurement of frequency deviations in one section and, in the other, for outputs at the mark and space frequencies to be used in checking and setting up the audio-frequency portions of the equipment.

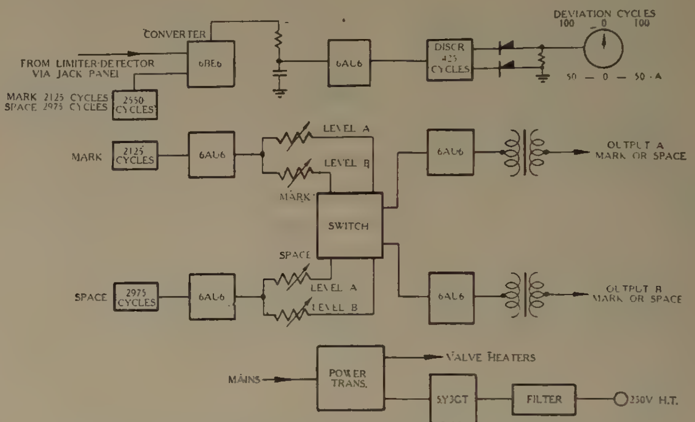
CONVERTER
FROM LIMITER-DETECTOR VIA JACK PANEL
6BE6
MARK 2125 CYCLES
SPACE 2975 CYCLES
2550 CYCLES
6AL6
DISCR 425 CYCLES
DEVIATION CYCLES
100 — 0 — 100
50 — 0 — 50 · A
MARK
2125 CYCLES
6AL6
LEVEL A
LEVEL B
MARK
SWITCH
SPACE
6AL6
OUTPUT A MARK OR SPACE
SPACE
2975 CYCLES
6AL6
LEVEL A
LEVEL B
6AL6
OUTPUT B MARK OR SPACE
VALVE HEATERS
MAINS
POWER TRANS.
5Y3GT
FILTER
250V H.T.

Fig. 10—Block schematic, frequency deviation indicator and voice frequency keying unit.

The deviation indicator derives its input signal from the output of the limiting amplifier in the limiter-detector unit, via the jack field, where the input circuit is normalled to channel *A* but can be patched to channel *B*, if desired. The first stage is a frequency converter employing a 6BE6 valve with the oscillator section arranged to operate at 2,550 cycles. If both mark and space signals are at their nominal frequencies, 2,125 and 2,975 cycles, respectively, the output of the converter stage derived from either will be at 425 cycles. A low-pass filter is employed to reduce spurious components at other frequencies, and the output from this filter is applied to an amplifier stage and then to a discriminator (Seeley-Foster type) with a dc center-zero meter in its rectifier circuit. The discriminator is arranged to have a reasonably linear characteristic over a range in excess of 100 cycles above and below a center-frequency of 425 cycles and the meter is calibrated over a 100-cycle range on each side of zero.

The deviation indicator is useful both in setting up a circuit and in checking operation during actual use. If the receiver is correctly tuned and the deviation of the transmitter is also correct, both mark and space indications will be zero. If an indication of 100 cycles, in opposite directions, is given for mark and space signals, this shows that the receiver is not correctly tuned and adjustment may then be made. If the mark signal reading is zero and the space reading 50 cycles, the indication is that the transmitter deviation is incorrect by that amount.

The vf keying unit comprises two oscillators, at 2,125 and 2,975 cycles, controls for level adjustment, a keyswitch for selecting mark or space frequencies, and two amplifier stages, each delivering output to appropriate points in the jack field. Accurate adjustments of level may be made by using the level indicator in the meter panel, and fixed attenuator pads also provided in the jack field permit a wide range of levels to be applied to various points in the equipment. Tests may be made for correct operation of the limiter-detector units, including the action of noise-suppressor and mark-hold circuits, relay unit, and frequency-deviation indicator.

## V. Performance Summary

### *Receiver*

The following is a summary of the typical performance characteristics of the receivers, operating with the main oscillator unit.

### *Sensitivity and Signal-to-Noise Ratio*

The figures were taken with input from signal generator modulated 30 per cent at 400 cycles through a 75-ohm resistor. Noise figures relate to the drop in output on removing modulation.

TABLE I

| *Frequency mc* | *Output Microvolts* | *Output Milliwatts* | *Noise db* |
|---|---|---|---|
| 3.15 | 1.0 | 10 | 12 |
| 5.05 | 0.5 | 10 | 9 |
| 7.2 | 0.5 | 10 | 7.5 |
| 14.3 | 1.0 | 10 | 10 |
| 27.0 | 1.0 | 10 | 9 |

*Image ratio*

| *mc* | *Image ratio db* |
|---|---|
| 3.15 | 100 |
| 5.05 | 73 |
| 7.2 | 76 |
| 14.3 | 60 |
| 27.0 | 38 |

*Selectivity* (measured on IF amplifier at 455-kc center frequency)

| Input ratio | | kc off resonance | | | | | |
|---|---|---|---|---|---|---|---|
| | | Wide | | Normal | | Crystal | |
| Times | db | + | − | + | − | + | − |
| 1.4 | 3 | 3.0 | 3.0 | 1.5 | 1.5 | 0.9 | 0.9 |
| 3 | 10 | 3.7 | 3.7 | 2 | 2 | 1.25 | 1.25 |
| 1,000 | 60 | 11 | 11 | 7.5 | 7.5 | 4.5 | 4.5 |

### *Automatic Gain Control*

At any frequency in the range the agc characteristic is within 4 db for input signals between 10 microvolts and 0.1 volt.

TABLE II

*Automatic Frequency Control*

| Control range (approximate) | | Resultant error in mark frequency |
|---|---|---|
| Within filter pass-band | Maximum | |
| ±300 cycles | ±2,000 cycles | Less than 60 cycles |

### *Tuning Ranges*

Radio-frequency, 1.5 to 30 mc in three ranges
- (a) 1.5 to 5 mc
- (b) 5 to 15 mc
- (c) 15 to 30 mc.

Beat oscillator, manual tuning ±5 kc.

High-frequency oscillator
- (a) Variable, 1.5–30 mc, as above.
- (b) Crystal range of 0.01 per cent of frequency by vernier control for each of three crystal positions.

### *Frequency Stability*

High-frequency oscillator crystals. For a variation in ambient temperature of ±20°C, the frequency change is not greater than 25 cycles for frequencies up to 15 mc and not greater than 50 cycles for frequencies above 15 mc.

Beat-frequency oscillator crystal. Within 0.01 per cent of 452.450 kc at any temperature in the range 10–60°C.

Variable-frequency hf oscillator. Better than 100 cycles per megacycle for temperature variations of ±5°C.

### *Limiter-Detector Unit*

The following data are typical performance figures for the important characteristics of the limiter-detector unit:

*Filter response.* Details are given in Fig. 6.

*Limiting amplifier.* Details are given in Fig. 8.

*Keying amplifier.* Output to relay coil (per channel) 15 mA on mark or space signals.

*Noise suppressor.* Removal of the signal from one channel reduces the relay current from this channel to less than 4 ma, and increases that of the other channel by not less than 90 per cent in an action time not greater than 2,000 microseconds.

*Mark-hold circuit.* Operates normally to within 3 db of the keying relay failing to follow signals caused by noise.

### *Relay Unit*

The rated characteristics of the relay unit are as follows:

*Normal keying speed*—50 bands (approximately 60 wpm on standard teletypewriter).

*Maximum recommended keying speed*—100 bauds.
*Polar output*—60-0-60 ma in 500 ohms.

*Frequency Deviation Indicator and VF Keying Unit*—
Performance figures are as follows:
*Deviation indicator range*—100-0-100 cycles, accuracy ±10 cycles.
*Accuracy of 2,550-cycle oscillator*—Within 10 cycles of nominal frequency.
*VF keying unit outputs*—Not less than ±10 dbm for each channel.
*Accuracy of 2,125-cycle and 2,975-cycle oscillators*—Within 10 cycles of nominal frequency.

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# Description and Evaluation of 100-Channel Distance-Measuring Equipment*

R. C. BORDEN†, C. C. TROUT†, MEMBER, IRE, AND E. C. WILLIAMS†

***Summary*—The common civil military program for navigational aids for use on and off the airways is based on a polar co-ordinate system from which a pilot may obtain his position relative to a known ground location. This navigational information is supplied in the form of azimuth and distance measurement from any preselected ground facility.**

**Under this program, azimuth information is supplied by the vhf omnirange which is now being installed on a country-wide basis. Development work on the final version of the other parameter of the navigation system, namely, the distance-measuring equipment, is nearly completed and plans are being made to add that equipment to the vhf omnirange facilities in the near future. Addition of the distance-measuring equipment to the instrument landing system (ILS) to provide continuous distance information to the runway touchdown point also is contemplated.**

**The distance-measuring equipment, commonly referred to as DME, employs pulse techniques and the interrogator-transpondor or challenge-reply principle so widely used in identification equipment during World War II. A portion of the frequency band, 960 to 1,215 mc, assigned to aids to air navigation, is being used for DME implementation. This report outlines the history and development of the present nationally adopted 100-channel DME, and describes the modifications necessary to convert earlier 50-channel equipment into equipment suitable for operating within a 100-channel system.**

**In operation, the aircraft unit of the DME, commonly called the interrogator, continuously transmits a train of short duration rf pulses. These pulses are received at the ground station, called the transpondor, and are shaped by suitable circuits into trigger pulses, which modulate the transmitter portion of the transpondor. The regenerated rf pulses transmitted by the transpondor, on a different frequency than the original received pulses, are in turn received by the airborne receiver section of the interrogator. Electrical measuring circuits within the interrogator measure the time elapsed between each pulse transmitted by the interrogator and the reception of its corresponding reply pulse from the transpondor. This time is directly related to the distance between the aircraft and the ground station and is converted into proper voltages for operating the pilot's distance indicator. The indicator is calibrated directly in miles. As many as 50 aircraft can use the same ground transpondor equipment simultaneously to obtain continuous distance information, without interference and with no reduction in performance.**

**The airborne interrogator must also be able to identify the ground transpondor, with which it is operating, as a check of the channel selecting mechanism. This is accomplished by the transmission of a third pulse, with a definite time relation to the distance measuring pulses from the transpondor. The interrogator recognizes the presence of this pulse and causes a 400-cps tone to be heard in the pilot's headset. With this arrangement several identification systems are possible, two of which are discussed.**

## I. INTRODUCTION

MANY ORGANIZATIONS have taken an active part in the development of distance-measuring equipment (DME). Among these are the Naval Research Laboratory, Canadian Research Council, Hazeltine Electronics Corporation, Federal Telecommunication Laboratories, U. S. Air Force, and the Civil Aeronautics Administration. The earliest operating DME was built by the Canadian Research Council in 1945, and was designed for operation in the frequency region of 200 mc. Early in 1946, an experimental DME designed for operation

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in the region of 1,000 mc was completed at the Naval Research Laboratory by the Combines Research Group. The primary purpose of constructing this model was to determine whether propagation in this frequency band would be suitable for DME use. With this question answered in the affirmative, the U. S. Air Force and the CAA awarded contracts to the Hazeltine Electronics Corporation and to the Federal Telecommunication Laboratories for the construction of development models of 1,000 mc distance-measuring equipment.

Upon delivery of these equipments, extensive laboratory and service tests were conducted by the Air Force and the CAA Technical Development and Evaluation Center.[1] The primary purpose of these tests was to evaluate the relative merits of two different types of DME channeling. At the time these equipments were ordered, the requirement existed that the DME system be capable of operation on any of 50 independent and noninterfering channels. It was intended to install a DME transpondor at each vhf omnirange[2] site to provide a complete azimuth-distance ground facility. There were to be 30 such omniranges with a 435-nautical mile square, each with a discrete frequency assignment, (these 30 frequencies would be repeated in each adjacent 500-mile square, noninterference being achieved by the geographical separation). In addition, DME transpondors were to be installed at each instrument landing site, there being 20 of these for each 435-nautical mile square, all with different frequency assignments. The means of obtaining the required 50 channels employed by Federal[3] and the means employed by Hazeltine[4] were different. The Federal equipment provided the 50 channels by assigning discrete radio frequencies for each channel, requiring a total of 100 radio frequencies with 2.5-mc separation. The Hazeltine approach to the problem consisted of selecting 26 radio frequencies, with adjacent spacing of 9.5 mc, within the assigned band and repeating each frequency four times. To prevent interference between channels on the same frequency, a system of double pulse transmissions, or pulse multiplex, was employed. In such a system, desired signals are selected and undesired signals rejected by proper design of video coding and decoding circuits.

The tests performed by the Air Forces and CAA were conducted under the auspices of RTCA subcommittee SC-40 (Standardization of DME Testing Procedures).[5] Both systems of channeling were found to be satisfactory, but the channeling problem became complicated further by a change in the operational requirement for the number of channels to be provided by the system. Re-examination of the over-all domestic short-range navigation requirements by RTCA subcommittee SC-22 led to the decision that a total of 100 channels must be provided rather than only 50.[6] Since separations of the order of 2.5 mc had been shown to be practical, and since the pulse-multiplex method of channeling had likewise proved its worth, the decision was made by SC-40 to combine the two techniques in order to obtain the 100 channels. Tests of the pulse-multiplex system indicated that the number of times an individual frequency could be employed might well be increased from four to ten; thus, 100 channels could be provided by only 20 radio frequencies (ten for interrogation and ten for reply). By employing an adjacent channel spacing of 2.5 mc, the entire DME system could be implemented within a total band of approximately 50 mc. Conservation of spectrum was held to be of extreme importance due to the requirements of RTCA subcommittee SC-31.[7] Among the requirements laid down by SC-31 was one prescribing that the 960 to 1,215-mc band must ultimately be used as a frequency medium for the entire short-range navigation system, including DME, omnirange, instrument landing, voice, and pictorial display.

A specification covering a system having the above channeling characteristics was prepared by SC-40. This specification was presented at the February, 1949, meetings of the International Civil Aviation Organization, and as a guide for DME development. Production equipment meeting the system requirements outlined by the SC-40 specification will be available in 1951. Detailed descriptions of the design and operation of 50-channel equipment of this type have been covered in literature,[1,3,4,8] and therefore, these details will not be repeated. This paper will be confined to those modifications required to expand the utility of the system to 100 operating channels.

## II. DESCRIPTION OF EQUIPMENT

Fig. 1 illustrates the basic operation of a DME system and indicates the proposed plan for installation of VOR transmitters and DME transpondors at the same site.

Figs. 2 and 3 are simple block diagrams of a basic DME interrogator and transpondor, respectively.

As a starting point, a summary of those equipment characteristics which differ in the two systems may be listed in Table I.

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[2] G. C. Hurley, S. R. Anderson, and H. F. Keary, "The CAA VHF Omnirange," TD Report.

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[6] RTCA Paper 76-48/DO-17, Report of SC-22 (Pairing of localizer, glide slope VHF omnirange, and DME frequencies); August 2, 1948.

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[8] J. Wesley Leas, "Some operational aspects of distance measuring equipment in the transition air navigation system," *Aero. Eng. Rev.*, vol. 9, pp. 31–36; January, 1950.

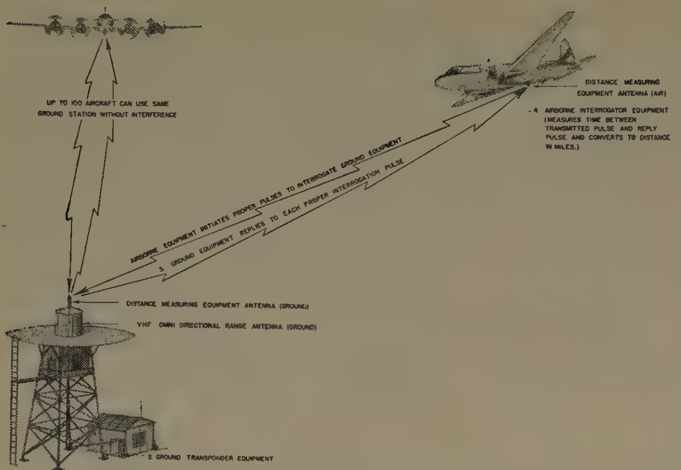
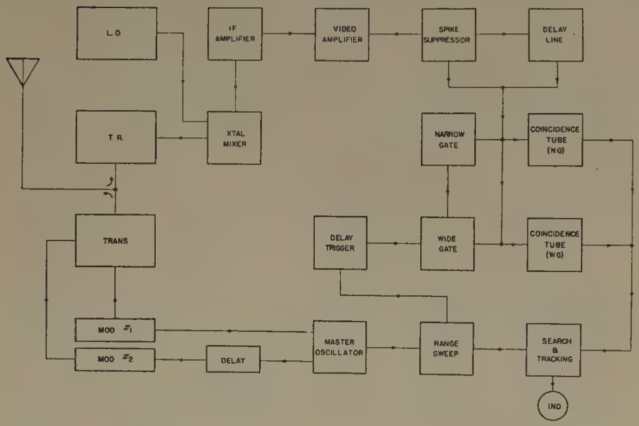

UP TO 100 AIRCRAFT CAN USE SAME GROUND STATION WITHOUT INTERFERENCE
DISTANCE MEASURING EQUIPMENT ANTENNA (AIR)
4 AIRBORNE INTERROGATOR EQUIPMENT (MEASURES TIME BETWEEN TRANSMITTED PULSE AND REPLY PULSE AND CONVERTS TO DISTANCE IN MILES.)
AIRBORNE EQUIPMENT INITIATES PROPER PULSES TO INTERROGATE GROUND EQUIPMENT
3 GROUND EQUIPMENT REPLIES TO EACH PROPER INTERROGATION PULSE
DISTANCE MEASURING EQUIPMENT ANTENNA (GROUND)
VHF OMNI DIRECTIONAL RANGE ANTENNA (GROUND)
2 GROUND TRANSPONDER EQUIPMENT

Fig. 1—DME system operation.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NARROW GATE
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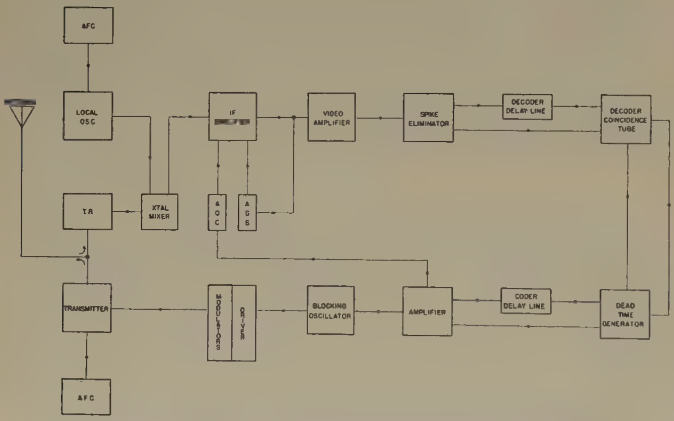
TRANS
DELAY TRIGGER
WIDE GATE
COINCIDENCE TUBE (WG)
MOD #1
MOD #2
DELAY
MASTER OSCILLATOR
RANGE SWEEP
SEARCH & TRACKING
IND

Fig. 2—Block diagram of the DME interrogator (airborne).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DECODER DELAY LINE
DECODER COINCIDENCE TUBE
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A G S
TRANSMITTER
MODULATORS
DRIVER
BLOCKING OSCILLATOR
AMPLIFIER
CODER DELAY LINE
DEAD TIME GENERATOR
AFC

Fig. 3—Block diagram of the DME transpondor (ground).

## III. Pulse Duration

Two methods of obtaining adjacent channel selectivity of the order of 70 db have been employed successfully in DME receivers. One of these, the spike-suppression circuit[1], requires the use of a wider pulse if the bandwidth is to be decreased. 2.5 µsec is sufficiently wide to permit satisfactory adjacent channel rejection with 2.5-mc channel separation. A typical response curve for a receiver employing spike suppression is shown as Fig. 4. Widening of the radio-frequency pulse from 1.5 to 2.5 µsec does not adversely affect the system in any other way, with the exception of the accompanying increase in duty cycle.

TABLE I

| | 50-channel | 100-channel |
|---|---|---|
| 1. Pulse duration | 1.5 µsec | 2.5 µsec |
| 2. Pulse spacings | 10, 15, 20, and 25* µsec or none† | 14, 21, 28, 35, 42, 49 56, 963, 70, 77 µsec |
| 3. Reply delay | 75 µsec | 115 µsec |
| 4. Reply frequencies | One of 13* frequencies, or one of 51† frequencies in 1087.5-1,215-mc band | One of 10 frequencies in 1188.5- to 1211.0-mc band |
| 5. Interrogation frequencies | One of 13* frequencies or one of 51† frequencies in 960- to 1087.5-mc band | One of 10 frequencies in 963.5- to 986.0-mc band |
| 6. Identification | Gap coding | Third pulse |
| 7. Traffic handling requirements | ——— | More stringent than 50-channel |

* Hazeltine.
† Federal.

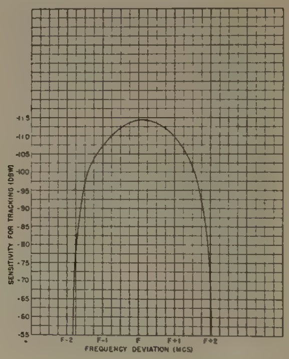

Fig. 4—Receiver bandwidth HEC interrogator (after spike suppression).

## IV. Pulse-Spacings

The requirement for additional pulse-spacing in order to provide ten modes has necessarily resulted in a requirement for longer magnetostriction delay lines. These longer lines have presented no serious problems inasmuch as the attenuation with length of such a line is practically negligible. These ten pulse-spacings have been paired for interrogation and reply to form ten standard modes, as shown in Table II.

TABLE II

| *Mode* | *Interrogation Spacing* | *Reply Spacing* |
|---|---|---|
| A | 14 | 17 |
| B | 21 | 70 |
| C | 28 | 63 |
| D | 35 | 56 |
| E | 42 | 49 |
| F | 49 | 42 |
| G | 56 | 35 |
| H | 63 | 28 |
| I | 70 | 21 |
| J | 77 | 14 |

The performance of coders and decoders at the longer pulse spacings has been completely satisfactory.

## V. Reply Delay

The increase in the reply delay from 75 to 115 μsec was required in order to accommodate the longer pulse-spacings employed in the 100-channel system. In the DME system, distance is measured by the interrogator as the time elapsed between the second interrogation pulse and the second reply pulse. Fig. 5(a) indicates the time relationship between these pulses when the aircraft is zero miles from a transpondor operating on Mode A. For a transpondor operating on Mode B, the delay line coil controlling the time of the first reply pulse, $r$ is simply moved seven microseconds closer to $r_2$, leaving the time interval between $i_2$ (the second interrogation pulse) and $r_2$ unchanged at 115 μsec. The additional 38 μsec are necessary to meet a requirement imposed when DME is used for instrument landing and the transpondor is not located at the touch-down point (where it is desirable for the aircraft distance indicator to display zero miles). By decreasing the time interval, $i_2$ to $r_2$, without changing the reply spacing, $r_1$ to $r_2$, this 38-μsec tolerance permits creation of an artificial zero up to approximately three miles from the transpondor. Fig. 5(b) shows the time relationship existent when the transpondor (still Mode A) is located approximately two nautical miles from the touch-down point. Note that the reply delay is adjusted for 90 μsec instead of 115 μsec. The 25-μsec difference is equivalent to a distance change of approximately two nautical miles.

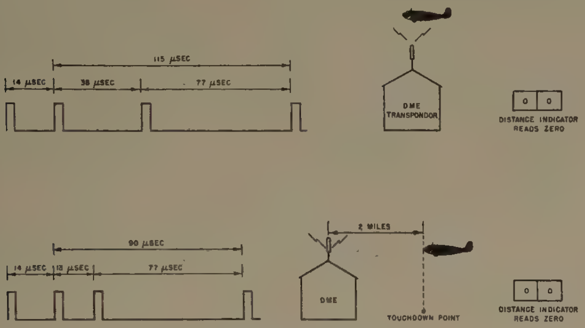

Fig. 5—Reply delay.

## VI. Interrogation and Reply Frequencies

In the case of the Federal DME, the frequency stability of the 50-channel units was sufficient to meet the 100-channel requirements; however, a considerable improvement in the reliability of the airborne transmitter AFC has been achieved in the 100-channel models. In order to meet the new channeling requirements, however, it was necessary to provide for cross-banding of the ten frequency pairs, whereas this was not required of the earlier 50-channel equipment. The total number of crystals required was nonetheless reduced and the pre-selector tracking problem was somewhat simplified due to the decreased receiver bandspread (25 mc as opposed to 127.5 mc).

In the case of the Hazeltine equipment, addition of an automatic-frequency-control system for the airborne equipment was required. This circuitry is similar to that employed for the ground transmitter of the 50-channel equipment. Airborne receiver stability was achieved through use of a single crystal, constant local oscillator frequency, and a variably tuned IF strip.

A contract has recently been awarded by the Civil Aeronautics Administration for a total of 450 100-channel DME transpondors. These transpondors will incorporate directly crystal-controlled transmitters. The output transmitter tube will be a type A-2328 triode operating as a power amplifier. Extensive tests performed at the manufacturer's plant have proven this tube to be a suitable power amplifier at the frequencies involved.

## VII. Identification

Earlier DME systems employed gap-coding as a means of transpondor identification. Gap-coding provides an indentification signal in the aircraft by virtue of periodic interruption of the reply pulses (performed by a keyer unit located at the transpondor) in accordance with characters of the Morse code. Interruption of the reply in this fashion may be detected in the aircraft either visually or aurally. Under conditions of very low traffic density (approximately ten aircraft per transpondor), this system is adequate. As the traffic loading increases, two factors tend to break down this identification system. As the result of transpondor countdown, the code indications in the aircraft are likely to be interrupted, and as a result of high random pulse levels, the code indicating device in the aircraft is likely to register spurious signals.

For reliable operation in a pulse-multiplex system with ten transpondors operating on each of the ten reply frequencies, an improved means of identification was dictated. The system selected involves the periodic transmission of a three-pulse reply signal from the transpondor. This third pulse will always have a definite time relationship (10.5 μsec later) with respect to the second reply pulse. A special gate in the interrogator will recognize the absence or presence of this coding pulse. The coding circuit, when activated by the third pulse, will result in the insertion of a 400-cps tone (generated within the interrogator) into the pilot's headset. Through this same headset the pilot will also be receiving both a voice and a Morse code identification from the VOR to which he is tuned. By employing the same mechanical keyer to activate the third pulse reply at the transpondor as is used to key the VOR identification, a synchronous relationship between the VOR code and the 400-cps tone is established. There have been several proposed systems for utilizing a synchronized 400-cps tone generated within the airborne DME to indicate to the pilot that he is tuned to the ground transpondor associated with the vhf omnirange or instrument landing system facility being used.

In one system the mechanical keyer is arranged to initiate transmission of the third pulse immediately following the last character of the VOR Morse code. Simultaneously, a long dash will be transmitted (at 1,020 cps) on the VOR channel which will have a duration equal to the period of transmission of the third pulse (approximately 1.5 seconds). The result, as heard in the pilot's headset, will be a beat note between 400 and 1,020 cps. Should the pilot inadvertently tune to a DME transpondor not associated with the VOR whose identification he is hearing, he will not hear this beat note, but will hear separate and distinct 400- and 1,020-cps long dashes. This is assured by asynchronous operation of keyer motors at the different VOR sites. The proper time relationship between the 400- and 1,020-cps tones is indicated in Fig. 6 under System I, as well as a typical example of improper time relationship due to incorrect channel selection by either the VOR or DME airborne units.

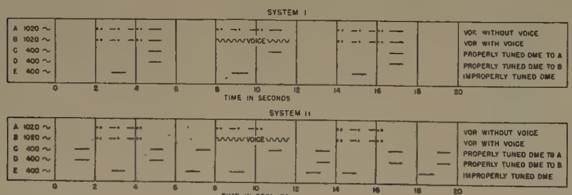

Fig. 6—VOR/DME identification system.

In another proposed system the transmission of the third pulse would be initiated for a period of about three-fourths second both before and after the transmission of the three Morse code characters from the VOR. The long dash would be eliminated from the VOR transmissions. This would result in the pilot hearing a 400-cps dash, followed by the VOR identification code at 1,020-cps and another 400-cps dash. In the event that improper channel selection should occur, the pilot will hear either one or the other of the 400-cps tones interfering with the VOR identification code letters. This system is illustrated in Fig. 6 under System II. It should be noted that this system does not cause any changes or additions to the VOR identification signals. Furthermore, such a signal, the long 1,020-cps dash, as heard in aircraft equipped with VOR only, may easily be confused with the letter "T."

On theoretical grounds, implementation of the 100-channel system, using only tested techniques (techniques employed in either or both of the 50-channel systems) presents no problems. Past experience, however, has indicated that until equipment meeting specific requirements has actually been built and operationally tested, it is sometimes difficult to predict performance with accuracy. For this reason, the placement of contracts for 100-channel equipment on a production basis was delayed until several developmental models became available. It was believed that satisfactory performance of such models under conditions which are to be encountered by future production equipment would be an excellent criterion of the soundness of 100-channel principles. Equipment manufacturers were urged to expedite delivery of 100-channel developmental models (both transpondors and interrogators), and simultaneously the Civil Aeronautics Administration initiated a project for combining two of the 50-channel interrogators (one pulse-multiplex and one high-stability type) in such a manner as to check various 100-channel characteristics. In addition to laboratory tests of 100-channel equipment, it was deemed essential to perform flight tests of the equipment under conditions of traffic density equivalent to that predicted for the foreseeable future.

Since the distance-measuring equipment, being an integral part of the Common Air Navigation and Traffic Control System defined by SC-31 of the RTCA, and now being developed under the direction of the Air Navigation Development Board, is a facility to be used by military as well as civil aircraft, a committee embracing both civil and military members was set up to direct the 100-channel DME evaluation program. This committee met on September 15, 1949, at the CAA Technical Development and Evaluation Center to discuss the various aspects of the test program. The test program was divided into four phases:

Phase I. Tests of the CAA combined pulse multiplex high stability 50-channel units.

Phase II. Laboratory tests of preproduction 100-channel units.

Phase III. Flight tests to determine the traffic handling capacity of 100-channel equipment.

Phase IV. Laboratory tests to determine the traffic handling capacity of 100-channel equipment.

Phases I and II consisted of rechecking equipment characteristics which are common to both 50- and 100-channel systems, and particularly the testing of combinations of 50-channel techniques required for implementation of the 100-channel system.[9]

Phases III and IV are considered of major significance inasmuch as they were set up for the purpose of providing more realistic justification that the 100-channel system, as designed, is capable of operating under the high traffic densities which ultimately are anticipated. Prior to the initiation of this test program, there was no suitable laboratory test equipment for determining traffic handling capacities of such system accurately, nor were there a sufficient number of ground stations geographically located so that an operational test could be arranged. Data of the type desired were previously limited to statistical analyses and to laboratory tests conducted with equipment not completely capable of simulating actual operating conditions. Although the data from these two sources were in reasonable accord, it was believed highly desirable to perform tests under more realistic conditions. If 100 transpondors were used in a 435 nautical square mile area, then there would be a total of ten transpondors transmitting on a common frequency, i.e., each transpondor frequency must be repeated ten times. It is conceivable that each of these transpondors may, at a given time, be supplying service to as many as 50 aircraft. In such cases, all of the transmitted replies will be eligible for reception by all of the 500 aircraft involved, subject to line-of-sight propagation restrictions. Any given aircraft will receive only one of the ten transpondors with a proper reply spacing, but all ten will be receved at the correct frequency. Interlacing of the replies from the nine improperly moded transpondors will result in an appreciable number of artificial replies having the correct mode being produced. In this case, the ability of the aircraft decoding circuits to sort out the replies of proper spacing and reject all others is extremely important. Furthermore, the ability of the searching and tracking circuits of a particular interrogator to recognize and respond only to those properly moded replies initiated by its own transmitter also is imperative. Inserting the line-of-sight limitations and a more realistic picture of traffic distribution, SC-40 concluded that the ability of the system to operate properly in the face of a loading of 20 aircraft for each of the nine improperly moded transpondors, and 50 aircraft for the properly moded transpondor, would guarantee satisfactory system operation for many years to come. The purpose of Phases III and IV of this program was to determine whether or not this ability was inherent in equipment which can be built now. Only Phase III will be herein discussed, Phase IV activity still being in progress.

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[9] TD Report No. 119.
R. C. Borden, C. C. Trout, and E. C. Williams, "Evaluation of 100-channel DME."

## VIII. Phase III

Performance of Phase III of the test program necessarily was delayed until ten transpondors had been modified to meet 100-channel requirements. It was intended originally to employ this number of transpondors in the test, distributed as follows: Indianapolis, Ind., and vicinity, (4); Lafayette, Ind., (1); Terre Haute, Ind., (1); Dayton, Ohio, (1); and Wilmington, Ohio, (3). All of the transpondors were operated at the same reply frequency (1,201 mc), but with different reply spacings.

Due to transfer of the Air Force's All-Weather Flying Division activities from Wilmington, Ohio, to Wright-Patterson Air Force Base at Dayton, Ohio, it was not possible for that group to provide all of the transpondors assigned to them for the test; however, one of three was made available. In order to avoid dependence upon air-to-ground communication to eight different sites during progress of the test flights, special schedules were set up based on time synchronization. Flight tests were conducted on November 29, 1949, and repeated on the following day.

In order to assure that the test aircraft was within propagation range of all transpondor sites, a flight test area was selected which was within 100 nautical miles of all sites. This area is indicated on the map, Fig. 7, and was roughly 35 nautical miles east of the Center. In order to assure further that strong signals would be received from all transpondors, all flights were made at altitudes above 8,000 feet indicated (mean elevation above sea level of all transpondors was approximately 800 feet). In tabulating the flight test data, unless a synchronized reply was received from a given transpondor, the "fruit" from that transpondor was not considered.

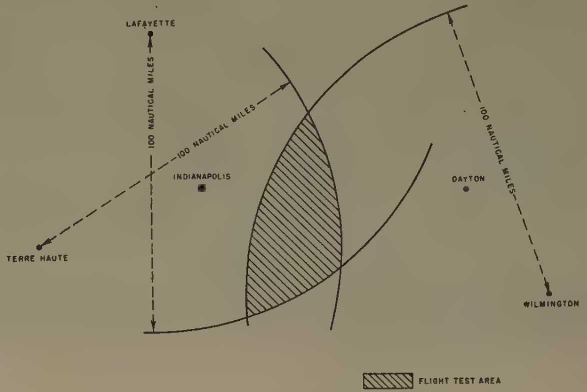

Fig. 7—Map of flight area.

Squittering of the transpondors was effected by turning up the automatic-gain-stabilization control while observing the transmitter current. A curve showing the relation of transmitter current and other transpondor characteristics to squitter rate is shown as Fig. 8. The squittering rates employed, ignoring the effect of count down, are roughly equivalent to the interrogation of each transpondor by 10, 20, 30, 40, and 50 aircraft, taken chronologically.

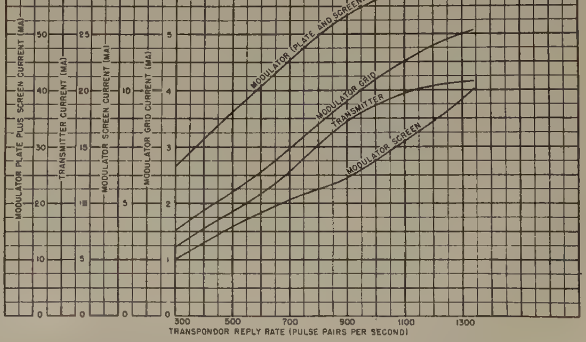

Fig. 8—Variation of transpondor parameters with reply repetition rate.

Two aircraft were employed during the first day's test, one from the Naval Air Test Center, Patuxent River, Md., and the other a CAA airplane N-181 based at the Technical Development and Evaluation Center. The test setup as installed on the Navy DC-4 airplane is shown in Fig. 9. Both test aircraft carried 100-channel airborne equipment. On returning to the ground, the flight engineers of both aircraft reported failure to receive replies from the Wilmington transpondor at all times during the flight. Replies were obtained from the remaining seven transpondors at all times.

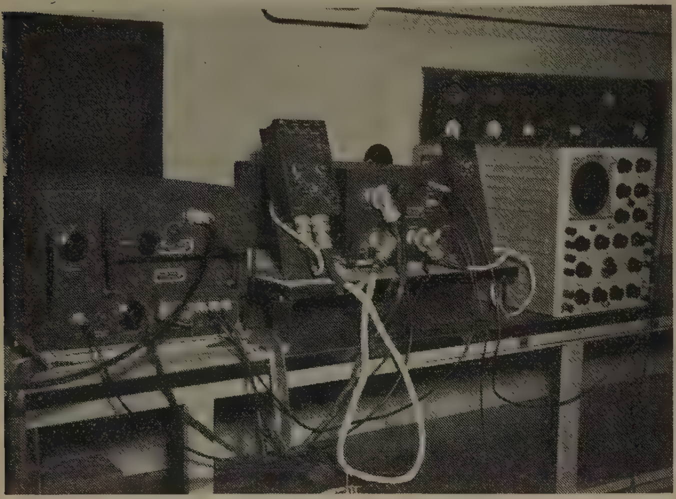

Fig. 9—Hazeltine 100-channel DME interrogator.

In the case of the Lafayette and Brownsburg, Ind., (Indianapolis area) transpondors, although strong synchronized replies were observed at the 1,500-pulse pair fruit level, it was not possible to "lock-on" the station. Examination of the oscilloscope carried in each airplane indicated that this failure was due to erratic transmission of the second reply pulse at high transmitter duty cycles. This is not a traffic handling capacity problem, but was a temporary malperformance of the transpondor. The fact that proper locking-on to the remaining five beacons was accomplished is evidence that excessive fruit was not responsible for this malperformance. It is significant that the Terre Haute, Ind., transpondor, farthest from the operating area, was locked-on at all fruit levels without difficulty. Furthermore, the Terre Haute transpondor was transmitting three pulses (two for reply and one for identification) at all times, thus adding to the total fruit level.

At no time during either flight was the search time of the airborne units appreciably lengthened, nor was the memory time of either unit extended. It is these two functions which generally fail first, under conditions of extremely high random pulse levels, i.e., fruit. One of the interrogators was adjusted for a 15-second memory time throughout the test.

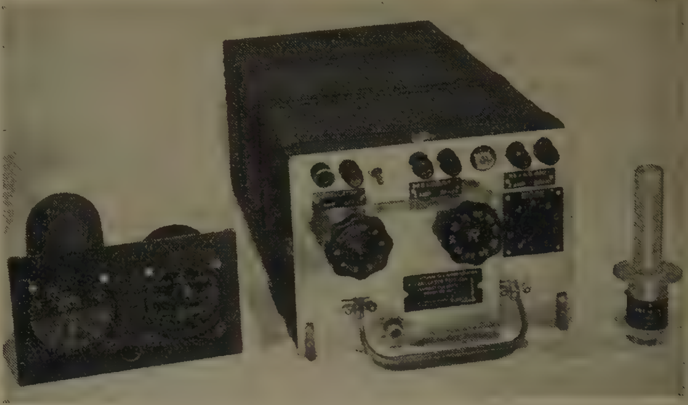

Fig. 10—Federal 100-channel DME interrogator.

These findings support the statistical figures which had been previously presented. In view of the fact that the airborne equipments demonstrated the ability to operate through the highest levels of fruit reached during the tests, the upper limit cannot be stated. On the other hand, since the maximum fruit level produced during the period of the flight test was in excess of that demanded of the equipment by current specifications, there is no reason to believe that the equipment will prove other than completely satisfactory from a traffic handling point of view.

In order to ascertain that the results of November 29 were not representative of some intangible but favorable condition, the tests were repeated on the following day. On November 30, both aircraft repeated the flights of the preceding day with minor schedule changes. The results obtained were the same.

## IX. Conclusions

The following conclusions may be drawn on the basis of these investigations of the transition from a 50-channel to a 100-channel DME system:

(1) The combination of 50-channel techniques to provide 100-channel equipment has been accomplished without detriment to the system.

(2) The traffic-handling capacity of present 100-channel equipment is adequate to accommodate any foreseeable future traffic densities.

(3) By compressing DME service into a relatively small portion of the total frequency spectrum allocated for air navigation aids, the way has been left open for system designers to employ considerable latitude in the design of future air navigation facilities complementary to the DME.

# Single-Conductor Surface-Wave Transmission Lines*

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***Summary*—This paper gives general information on single-conductor transmission lines, in particular on lines having a dielectric-coated conductor. A design chart is included and experimental results are presented.**

## I. Introduction

THE TOTAL FIELD which is excited on any waveguide consists of two parts: the field of the regular wave and a supplementary field which takes care of the continuity of the field at places where the guide is excited, or, where its geometrical structure deviates from the ideal one. In the case of closed waveguides with only one propagating mode, the supplementary field is made up of nonpropagating wave modes with purely reactive power flow. On open guides the supplementary field is a radiation field. Consequently, the excitation of an open guide is connected with more or less radiation loss, and any kind of distortion of the field results in radiation loss. From this point of view the open guides appear inferior to the closed ones. However, the open guides have two advantages: they are less bulky and less expensive; and, since the field is not confined to a small area, the field density is low. This, in return, has the effect that, under comparable conditions, the resistive and dielectric losses of open guides are much smaller than those of closed guides.

The only open waveguide which has been thoroughly investigated so far is the two-wire line. No doubt, this line has a range of applications in which it is superior to cables because of low losses and lower costs. The dielectric guide, known since 1910,[1] received more interest after the dielectric rod antenna was discovered. Although a dielectric wire is capable of propagating a wave mode of very low attenuation[2,3] (the so-called "dipole mode"), its application for transmission lines is not very promising for mechanical reasons, and because the extension of the field is very large. This paper deals with the third type of open waveguides, namely, the single-conductor guide which shows very promising aspects with regard to practical application.

The single conductor as a means for propagating a nonradiating wave is of particular interest—for practical reasons since it appears to be the simplest waveguide; for theoretical reasons since it occupies a unique place among waveguides in that it would not be a guide at all under ideal conditions, that is, if the conductivity were infinite and the surface perfectly clean and smooth. This fact already indicates that an ordinary copper wire, with its high conductivity and rather smooth surface, is not suitable as a waveguide, except for extremely high frequencies where the effect of the finite conductivity becomes more apparent. Although it has been known for more than 50 years, from a paper by Sommerfeld[4], that a conductor of finite conductivity should guide a nonradiating wave mode, actually this wave mode has not been utilized for transmission lines. There are antennas with single-wire feeds, but, the conditions are such that the energy is transported by a radiating wave like on long wire antennas rather than by Sommerfeld's surface wave. In principle, this wave should appear on every antenna built from wires. Experience shows, however, that the field of such antennas is practically the same as the field of ideally conducting antennas and only slightly modified by the finite conductivity. This modification has nothing to do with Sommerfeld's wave, which is an independent solution of Maxwell's equations. The reason why Sommerfeld's wave does not enter into antenna problems is that the usual coupling devices excite chiefly the supplementary field, i.e., the radiating wave. As shown later, Sommerfeld's wave requires launching devices of very large dimensions, and this is because the field extends very far from the conductor; in other words, the field decreases slowly in the radial direction. The large extension of the field makes Sommerfeld's wave impractical not only because of the difficulties in exciting the wave, but also for the reason that a large clearance around the conductor is necessary in order to avoid severe distortion of the field. Furthermore, the large extension of the field makes the wave very sensitive to small bends or sags in the line. The wave would become practical only for frequencies above 10,000 mc.

It is of interest that the usual imperfection of the surface of a conductor can be sufficient to convert Sommerfeld's wave into another wave mode, a wave mode which is mainly determined by the surface conditions. This type of wave is of particular interest since the extension of the field can be controlled by modifications of the conductor surface. It is this type of wave which is considered in this paper.

## II. Field of the Wave

The only wave mode which is guided by a single conductor is a radially symmetrical transverse magnetic mode having in cylinder co-ordinates the field components[5]

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† Signal Corps Engineering Laboratories, Fort Monmouth, N. J.

[1] D. Hondros and P. Debye, "Elektromagnetische wellen an dielektrischen drähten," *Ann. Phys.*, vol. 32, p. 465; 1910.

[2] C. H. Chandler, "An investigation of dielectric rod as wave guide," *Jour. Appl. Phys.*, vol. 20, p. 1188; December, 1949.

[3] W. M. Elsasser, "Attenuation in a dielectric circular rod," *Jour. Appl. Phys.*, vol. 20, p. 1193; December, 1949.

[4] A. Sommerfeld, "Fortpflanzung elektrodynamischer wellen an einem zylindrischen leiter," *Ann. Phys. u. Chemie*, vol. 67, p. 233; December, 1899. (See J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., p. 527; 1941.

[5] Mathematically there exist also wave modes with cyclic periodicity around the cylinder (investigated by D. Hondros, "Uber elektromagnetische drahtwellen," *Ann. Phys.*, vol. 30, p. 905; 1909) which are damped out extremely fast and practically not excited.

$$E_r = jAhH_1(\gamma r)e^{j(\omega t - hz)}$$

$$E_z = A\gamma H_0(\gamma r)e^{j(\omega t - hz)} \qquad \text{with} \qquad \gamma = \sqrt{k^2 - h^2}$$

$$H_\phi = jA\sqrt{\frac{\epsilon}{\mu}}\, kH_1(\gamma r)e^{j(\omega t - hz)} \qquad k = \frac{2\pi}{\lambda};$$

($\lambda$=free-space wavelength, $h$=propagation constant). $H_0$ and $H_1$ are Hankel functions of the first or second kind, depending on which root of $\gamma$ is chosen. The parameter $\gamma$ is a measure of the steepness of the field decrease in the radial direction. It also determines the reduction in phase velocity. If we call the diameter of a circle around the wire, within which a certain percentage (for instance, 90 per cent) of the energy is propagated, the diameter of the field, then we may say: $\gamma$ determines the diameter or the extension of the field. The larger $\gamma$, the smaller the diameter of the field and the smaller the phase velocity of the wave.

The boundary conditions at the surface of the guide require the continuity of $E_z$ and $H_\phi$. They can be satisfied, as shown by Sommerfeld,[4] on a conductor with finite conductivity. Since the ratio between the tangential electric and magnetic field components at the surface of a good conductor is very small, $\gamma$ is very small compared with $k$. Hence, the distance from the wire at which the field becomes negligible is a large number of wavelengths.

The boundary conditions can also be satisfied on conductors of infinite conductivity if the surface is coated with a dielectric or magnetic layer. This case, first considered by Harms,[6] is of particular interest. There, $\gamma$ becomes purely imaginary, provided the losses in the layer are negligible. Since the thickness of this layer can be varied, the value of $\gamma$ and therefore the extension of the field can be adjusted to any desired value. For frequencies above 100 mc, reasonable field diameters are obtainable under conditions for which the phase velocity of the surface waves is little smaller than the velocity of light and the energy propagated within the layer amounts to a small fraction of the total energy. These facts are illustrated by the curves in Fig. 1. They show the 90 per cent radius of the field ($\rho_{90\%}$), the reduction in phase velocity ($\delta v$), and the fraction of the energy propagated within the dielectric layer $N_i/N$, as functions of the thickness of this layer. The curves refer to a case where the wire radius is 0.1 cm and the frequency 3,000 mc. The dielectric layer is assumed to be enamel with a dielectric constant of 3. We see from the curves that very thin layers of dielectric already affect the extension of the field considerably while the phase velocity is reduced by less than 1 per cent. Since the part of the energy which is propagated within the layer is very small, the dielectric losses are almost negligible. It is of interest that within the frequency range above 100 mc a dielectric layer of $10^{-3}$ cm has more effect on the extension of the field than the conductivity of good conductors.

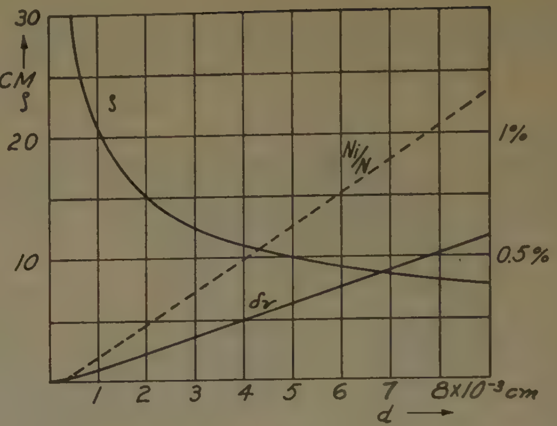

Fig. 1—Ninety-per cent power flow radius of the field $\rho_{90\%}$, reduction in phase velocity $\delta v$, and relative power propagated in the dielectric layer $d$ as functions of the thickness of the layer (wire radius 0.1 cm; $\epsilon_l/\epsilon=3$; $f$=3,000 mc).

It should be mentioned that the boundary conditions required for the existence of the surface wave can also be satisfied as average conditions on conductors with rough surfaces like threaded or twisted wires. If the corrugations become very large, like the disk-loaded wires used in some traveling-wave tubes, the surface wave degenerates to an accompanying field and is no longer the main carrier of the energy.[7] Therefore, the losses are greatly increased.

## III. EXCITATION OF THE WAVE

The excitation device for any surface wave should simulate the field distribution within a cross plane of the guide as well as possible since any deviation from this distribution is compensated by the supplementary field and thus results in a radiation loss. Theoretically, the simulation of the field could be achieved by a layer of electric or magnetic dipoles—in our case, for instance, by axially directed electric dipoles covering the surface of a metal plane which is perpendicular to the wire, with a density proportional to the longitudinal electric field component of the wave. Since the field of the wave extends without limit, the dipole distribution should also extend to infinity. Since in the actual case the field is simulated only within a finite area, a certain radiation loss is unavoidable.

A practical method of simulating the field of the surface wave on a single conductor becomes evident if we compare the field of this wave with the field of a wave in a coaxial line. Consider first an ideal coaxial line with infinite conductivity. The wave which is propagated in such a line is a transverse electric and magnetic mode. If the inner conductor is coated with a dielectric layer, the wave obtains a longitudinal electric component like that of the surface wave, and can be described mathematically as the superposition of two waves: one which has a similar field distribution to the surface wave (described by Hankel functions) and another, the field of which decreases in the *inward* direction (described by a

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[6] F. Harms, "Electromagnetische wellen an einem draht mit isolierender zylindrischer hülle," *Ann. Phys.*, vol. 23, p. 44; 1907.

[7] For a study of disk-loaded guides, see W. Rotman," A Study of Single Corrugated Guides," Air Force Cambridge Research Laboratories, Cambridge, Mass.; February, 1950.

Bessel function). The field of the latter may be considered as the image of that part of the first wave which would extend beyond the outer conductor of the line. If the diameter of this conductor becomes larger and larger, the amplitude of this field decreases more and more, and finally becomes negligible if the diameter is very large.

The effect of the diameter of the outer conductor on the power transmission is demonstrated by the curves in Fig. 2 which show the transmitted power $N$ as a function of the ratio between outer $b$ and inner $a$ radii under the condition that the current $I$ in the line is maintained constant. If we define $N/I^2$ as the impedance $Z$ of the line, the curves give a plot of $Z$ versus $b/a$ for an ordinary coaxial line, and another coaxial line, the inner conductor of which is coated with a thin dielectric layer. In both cases the losses are neglected. The broken curve which refers to the ordinary coaxial line shows that the impedance approaches infinity if the outer conductor becomes larger and larger. This indicates that the outer conductor is essential even if its radius is already very large. The solid curve which refers to the line with dielectric-coated inner conductor approaches a finite value if the diameter of the outside conductor of the line is increased. Therefore, the outer conductor has no appreciable effect on the wave if its radius is sufficiently large; and the surface-wave transmission line can be considered as a dielectric-coated coaxial line, the outer conductor of which has such a large diameter that it has no effect on the field of the wave. If we would have considered the finite conductivity, the curve for the ordinary coaxial line would also approach a finite limit, but for a much larger ratio of $b/a$. The finite conductivity creates a wave mode which can be guided by the inner conductor only, namely, Sommerfeld's wave.

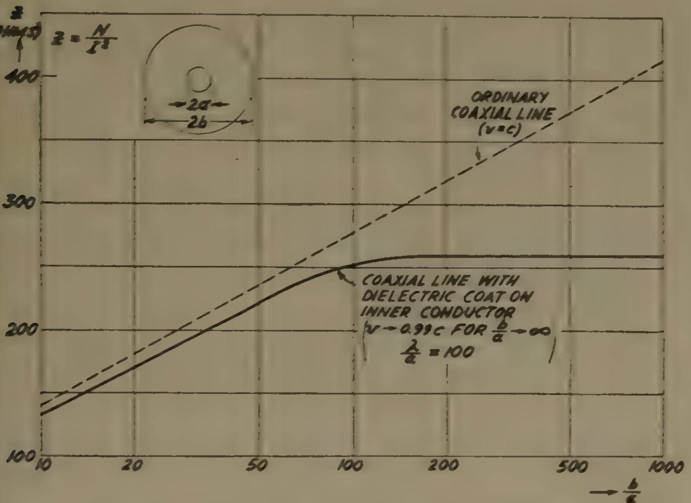

Fig. 2—Effect of the diameter of the outer conductor of a coaxial line on the power transmission if the inner conductor is coated with a dielectric layer.

The consideration of the surface waveguide as a coaxial line with a remote outer conductor immediately answers the question of how the surface wave can be launched (see Fig. 3.) Starting with a coaxial line, the inner conductor of which has a dielectric coat, the outer diameter is gradually increased until it is so large that it does not affect the field considerably. The field distribution then approaches that of the open line. Thus, as shown in Fig. 3, the wave can be launched by means of a metal cone which is connected to the outside conductor of a coaxial feed line. The inner conductor of that line is connected to the surface waveguide. The field lines shown in this figure give a rough idea of how the surface wave develops. The complete setup for a surface-wave transmission line is shown in Fig. 4.

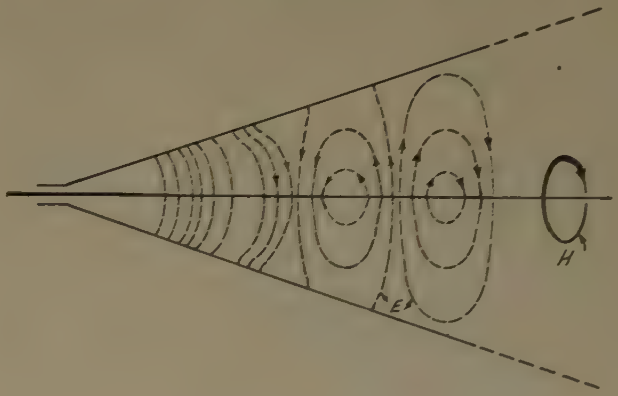

Fig. 3—Launching of the surface wave.

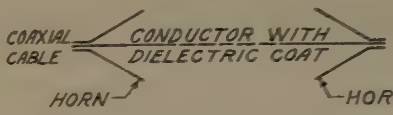

Fig. 4—Sketch of a surface-wave transmission line.

## IV. LOSSES OF A SURFACE-WAVE TRANSMISSION LINE

Losses of surface-wave transmission lines with a dielectric-coated wire consist of three parts: the conductivity loss in the wire, the loss in the dielectric coat, and the loss, effected by the launching device, due to a partial excitation of the radiating mode. In addition to these losses there are further losses, introduced by supports, bends, and sags, which shall be considered later. The conductivity loss and dielectric loss are proportional to the length of the line. They can be calculated from the field distribution of the wave. The launching loss is independent of the length of the line and can be determined with fair accuracy by the following consideration: At the receiving end, that portion of the wave energy will be received which travels within the area of the aperture of the horn. The wave energy outside this area will be lost. The ratio between the received energy and the total energy determines the efficiency of the receiving horn. Because of the reciprocity theorem the efficiency of the transmitting horn must be the same as the efficiency of the receiving horn.

Fig. 5 shows curves from which the three components of the loss can be determined. These curves are used for the actual design of transmission lines.[8] The formulas for the conductivity loss (for copper) and the dielectric loss are given in Fig. 5. $P$ and $Q$ are functions of a variable $G$, which, for thin dielectric layers, is proportional to the thickness of the layer. The curves $\rho'/a'$ give the required radius of the horns $\rho'$, measured in multiples of the guide radius $a'$ (conductor radius plus dielectric layer thickness) if a loss of 1, 2, 3, · · · db for both horns together is allowed.

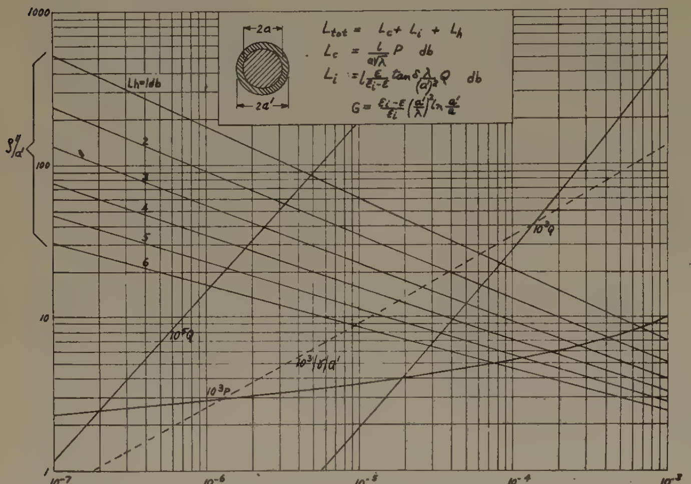

Fig. 5—Design curves ($a$, $a'$, $\rho'$, and $\lambda$ in cm, length $l$ of the line in feet).

The conductivity loss which is proportional to $P$ changes little with increasing layer thickness. The dielectric loss rises much faster (see curve $Q$); however, since only a small fraction of the energy is propagated in the layer, the dielectric loss is usually small compared with the conductivity loss even for rather thick layers. The diameter of the horns for a given launching loss, or the launching loss for a given size of the horns decreases with increasing thickness of the insulation since the field is more concentrated.

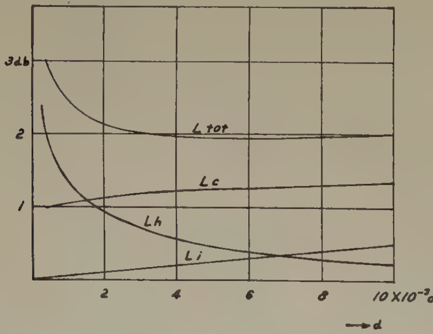

Fig. 6—Conductivity loss $L_c$, dielectric loss $L_i$, launching loss $L_h$, and total loss $L_{tot}$; as functions of the thickness $d$ of the dielectric layer. (Length of the line 120 feet, wire diameter 0.2 cm, horn diameter 33 cm, $f = 3{,}000$ mc, $\epsilon_i/\epsilon = 3$.)

A numerical example for the relation between the losses and the thickness of the dielectric layer is shown in Fig. 6. The curves are calculated for a copper wire of 0.2 cm diameter and 120-foot length. The dielectric layer is assumed to be enamel ($\epsilon_i/\epsilon = 3$; $\tan\delta = 8 \times 10^{-3}$). The opening of the horns is 33 cm and the frequency 3,000 mc. If the thickness of the layer is increased, the total loss of the line $L_{tot}$ first goes down because of the decreasing launching loss. After passing a flat minimum it rises since the dielectric losses become more and more influential.

A curve like this was verified experimentally. The measurements were made for a frequency of 1,600 mc and a wire diameter of 0.26 cm. The wire was outdoors for several months and had a rather thick corrosion layer. Under this condition the measured loss was 3 db. Then most of the corrosion was removed by sandpaper. The result was an increase of the loss to about 3.6 db due to the increased extension of the field. The expected loss for a wire with perfectly clean surface is much higher. Then several coats of polystyrene solution were applied by means of a brush. The loss went down more and more until a minimum of 1.7 db was reached. Further increase of the dielectric layer gave a slight increase of the loss. The thickness of the dielectric coat varied more than 1:10 along the wire; thus it was not possible to measure the average thickness. This test indicated that the wave is not sensitive to inhomogeneities of the guide.

The frequency response of a surface-wave transmission line is demonstrated in Fig. 7. In this case, an ordinary enamel wire of 0.2-cm diameter and $5 \times 10^{-3}$ cm enamel thickness ($\epsilon_i/\epsilon = 3$; $\tan\delta = 8 \times 10^{-3}$) is assumed. The length of the line is 120 feet and the opening of the horns 33 cm. The conductivity loss $L_c$ and the dielectric loss $L_i$ increase with frequency while the launching loss $L_h$ decreases. The total loss has a minimum of about 2 db. The actual minimum loss measured for this setup was 2.2 db.

Fig. 8 shows a comparison between the losses of coaxial cables, copper waveguides, and ordinary enamelled copper wires used for surface-wave transmission lines. The curves for the waveguides and the wires are calculated curves. The #12 wire (0.205-cm diameter) has an enamel coat of 0.00375-cm thickness and the #8 wire (0.325-cm diameter) a coat of 0.0025 cm. The dielectric

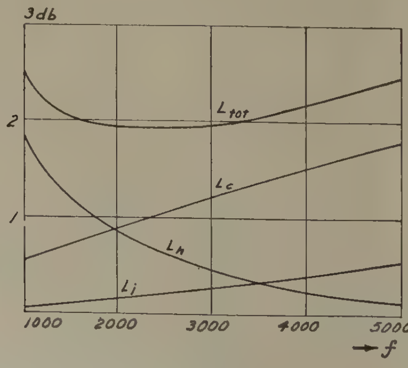

Fig. 7—Frequency dependence of the losses of a line. (Length, 120 feet; wire radius, 0.2 cm; dielectric layer thickness, 0.005 cm; $\epsilon_i/\epsilon = 3$, $\tan\delta = 8 \times 10^{-3}$, horn diameter, 33 cm.)

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[8] For the theory on which these curves are based, see G. Goubau, "Surface waves and their application to transmission lines," *Jour. Appl. Phys.*, vol. 21, p. 1119; November, 1950.

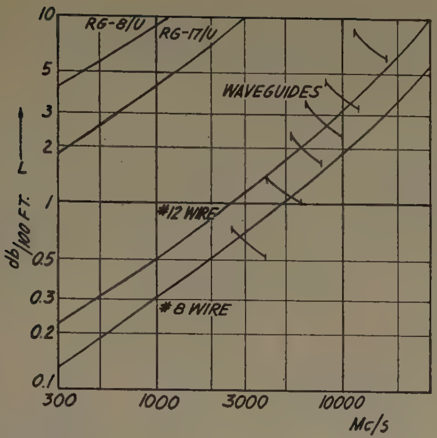

Fig. 8—Comparison between the losses of coaxial cables, copper waveguides, and ordinary enamel copper wires.

constant was assumed to be 3 and the power factor $8\times10^{-3}$. There is some uncertainty about the value of the power factor at very high frequencies; the value of $8\times10^{3}$ is valid at low frequencies. Therefore, a measurement on the #12 wire was performed at 23,000 mc, where the dielectric loss is comparable with the conductivity loss. The measured loss per 100 feet was 9.5 db while the curve shows 7.5 db. Assuming the discrepancy is due entirely to a larger power factor of the enamel, its value would have to be 80 per cent higher at this frequency range than for low frequencies.

Since there is practically no limitation on the diameter of the conductors as far as other modes are concerned, the loss of surface waveguides can be made much smaller than indicated in Fig. 8. Fig. 9, for instance, shows the theoretical loss of a line with a diameter of 1.42 cm and a polyethylene coat of 0.34 cm. There the loss is only a few db per mile.

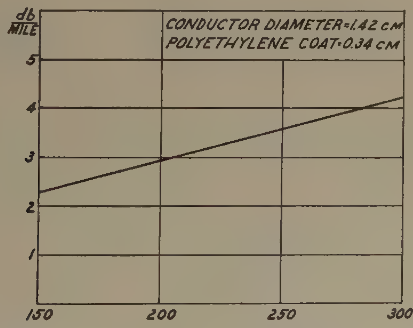

Fig. 9—Theoretical loss of a line for long-distance communication.

What has been considered so far is only the regular loss which is present under ideal conditions where the conductor is perfectly straight and the field is not disturbed by any object in the vicinity of the conductor. Actually the wire has sags and small bends, and it needs supports if the length exceeds a certain limit. Furthermore, since it is in the vicinity of the earth the field cannot extend without limits, as it does in the theory. All these effects cause additional loss by radiation or absorption, and the question is how does this additional loss compare with the regular loss. Although this question requires further studies, it can be answered in part. In our experiments, which were made with lines of enameled wires having a loss of the order of one db per 100 feet, the usual sag and small bends effected by supports caused no measurable increase of the loss. For instance, a test was made with a line of 600-foot length, stretched 4 to 8 feet above the ground along the slope of a hill. The wire was 0.32 cm in diameter and had an enamel coat of 0.0025 cm. The horns had a 33-cm diameter and the frequency was 1,600 mc. The theoretical loss was 4.5 db and the measured loss about 5 db. The number of supports, which consisted of waxed strings, was varied and there was no measurable effect on the loss observed. The situation could be different in cases like that of Fig. 9 where the line has a very low attenuation and a large number of supports are required within a stretch of 1 db loss. Although there are, as yet, no experimental results available for such lines, approximate calculations show that even in such cases the supports should not increase the loss considerably, provided they are properly made. The proximity of the ground has under reasonable conditions no noticeable effect on the performance of the line.

Another cause of losses should be mentioned, namely, rain and ice. Rain is troublesome only if it accumulates on the conductor in the form of drops. These drops act like small radiating dipoles. A water film is of no consequence in its effect on the performance of the line. It merely enlarges the thickness of the dielectric layer and behaves like a rather good dielectric because the ratio, power factor to dielectric constant minus one, which determines the losses, is very small. The radiation loss of drops can be calculated approximately. It decreases with the second power of the conductor radius and the fourth power of the wavelength. In the frequency range below 1,000 mc the loss should be negligible. For higher frequencies some observations were made. On the 600-foot line mentioned before (diameter 0.32 cm) an increase of the loss by 1.5 db was measured for a frequency of 1,600 mc under conditions where parts of the wire were densely covered with big drops. The same increase occurred under similar conditions at 3,300 mc on the 120-foot line (diameter 0.2 cm). In case of vertically inclined antenna feed lines, where drops cannot stay settled, the effect of rain is much less. For experimental purposes surface-wave transmission lines were installed in a television relay link which was operated over a period of several weeks. The frequency was 2,000 mc. The lines showed no noticeable increase of loss during rain or snow fall.

The formation of ice on the lines is serious since ice forms thick layers with rather poor dielectric properties. However, the formation of ice in most locations can be easily prevented by electric heating.

## V. ACKNOWLEDGMENTS

These experiments were made at the Signal Corps Engineering Laboratories. The author wishes to express his appreciation to John Hessel; R. E. Lacy, Chief and

Deputy Chief, Radio Communication Branch; J. J. Egli, Chief, Radio Relay & Microwave Section; and especially to the group who co-operated in the measurements, C. E. Sharp, L. R. Battersby, E. A. Conover, and, particularly, A. Meyerhoff, who also assisted in the preparation of the material.

# Standards on Pulses: Definitions of Terms—Part I, 1951*

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## I. INTRODUCTION

THE MEANINGS of commonly used terms in pulse work have often been a matter of disagreement. The IRE Standards Committee, faced with the fact that different technical committees proposed different definitions for the same term, and wishing to try to introduce uniformity where little has existed, set up a special task group with wide representation of special interests to propose standard definitions of terms concerned with pulses. Months of work by the task group and intensive critical review by the Standards Committee has led to results the first half of which are given below. These are necessarily compromises. The Standards Committee urges IRE members (a) to try to use the terms according to the definitions below, so that reasonable uniformity may be achieved, and (b) to take into account that in this particularly controversial region many compromises have been necessary so that favorite meanings and uses may appear not to have been considered, whereas in actuality it is unlikely that the very thorough review of the field has failed to unearth them for the Committee's consideration. The present set of definitions has been judged sufficiently important to print at this time; it is hoped that the second half will be ready for printing within nine months.

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Since many pulse shapes are possible, and a clear concept of the ones under discussion is desirable, it may be helpful to use drawings of pulse shapes, pulse times, magnitudes, durations, and the like, to show how these quantities apply.

In these definitions, linear superposition of pulses and possibly other waveforms is understood. Since it is possible to generate pulses whose characteristics may not seem to be adequately covered by the definition, it has been assumed that complex pulses can be analyzed in terms of more fundamental pulses and waveforms somewhat as a complex periodic wave can be considered as the sum of a fundamental and harmonics.

## II. DEFINITIONS

**Average Absolute Pulse Amplitude.** The average of the absolute value of the instantaneous amplitude taken over the pulse duration.

Note—By "absolute value" is meant the arithmetic value regardless of algebraic sign.

**Average Pulse Amplitude.** The average of the instantaneous amplitude taken over the pulse duration.

**Crest Factor of a Pulse.** The ratio of the peak-pulse amplitude to the rms pulse amplitude.

**Leading Edge Pulse Time.** The time at which the instantaneous amplitude first reaches a stated fraction of the peak pulse amplitude.

**Mean Pulse Time.** The arithmetic mean of the leading edge pulse time and the trailing edge pulse time.

Note—For some purposes the importance of a pulse is that it exists (or is large enough) at a particular instant of time. For such applications the important quantity is the *Mean Pulse Time.* The *Leading Edge Pulse Time* and *Trailing Edge Pulse Time* are significant primarily in that they may allow a certain tolerance in timing.

**Peak Pulse Amplitude.** The maximum absolute peak value of the pulse excluding those portions considered to be unwanted, such as spikes.

Note—Where such exclusions are made, it is desirable that the amplitude chosen be illustrated pictorially.

**Pulse.** A variation of a quantity whose value is normally constant; this variation is characterized by a rise and a decay, and has a finite duration.

Note 1—The word "pulse" normally refers to a variation in time; when the variation is in some other dimension, it shall be so specified, such as "space pulse."

Note 2—This definition is broad so that it covers almost any transient phenomenon. The only features common to all "pulses" are rise, finite duration, and decay. It is necessary that the rise, duration, and decay be of a quantity that is constant (not necessarily zero) for some time before the pulse and has the same constant value for some time afterwards. The quantity has a normally constant value and is perturbed during the pulse. No relative time scale can be assigned.

**Pulse Amplitude.** A general term indicating the magnitude of a pulse.

Note 1—For specific designation, adjectives such as average, instantaneous, peak, rms (effective), etc., should be used to indicate the particular meaning intended.

Note 2—Pulse Amplitude is measured with respect to the normally constant value unless otherwise stated.

**Pulse Amplitude, Average.** See *Average Pulse Amplitude.*

**Pulse Amplitude, Average Absolute.** See *Average Absolute Pulse Amplitude.*

**Pulse-Amplitude Modulation (PAM).** Amplitude modulation of a pulse carrier.

**Pulse Amplitude, Peak.** See *Peak Pulse Amplitude.*

**Pulse Amplitude, RMS (Effective).** See *RMS (Effective) Pulse Amplitude.*

**Pulse Bandwidth.** The smallest continuous frequency interval outside of which the amplitude of the spectrum does not exceed a prescribed fraction of the amplitude at a specified frequency.

Caution—This definition permits the spectrum amplitude to be less than the prescribed amplitude within the interval.

Note 1—Unless otherwise stated, the specified frequency is that at which the spectrum has its maximum amplitude.

Note 2—This term should really be "Pulse Spectrum Bandwidth" because it is the spectrum and not the pulse itself that has a bandwidth. However, usage has caused the contraction and for that reason the term has been accepted.

**Pulse Carrier.** A pulse train used as a carrier.

**Pulse Decay Time.** The interval between the instants at which the instantaneous amplitude last reaches specified upper and lower limits, namely, 90 per cent and 10 per cent of the peak-pulse amplitude unless otherwise stated.

**Pulse Duration.** The time interval between the first and last instants at which the instantaneous amplitude reaches a stated fraction of the peak pulse amplitude.

**Pulse-Duration Modulation (Pulse-Length Modulation) (Pulse-Width Modulation).** A form of pulse-time modulation in which the duration of a pulse is varied.

Note—The terms "pulse-width modulation" and "pulse-length modulation" are also used to designate this system of modulation but the term "pulse-duration modulation" is preferred.

**Pulse Duty Factor.** The ratio of the average pulse duration to the average pulse spacing.

Note—This is equivalent to the product of the average pulse duration and the pulse repetition rate.

**Pulse Frequency Spectrum.** See *Pulse Spectrum.*

**Pulse Interleaving.** A process in which pulses from two or more sources are combined in time-division multiplex for transmission over a common path.

**Pulse Interval.** See *Pulse Spacing.*

**Pulse-Interval Modulation.** A form of pulse-time modulation in which the pulse spacing is varied.

**Pulse-Length Modulation.** See *Pulse-Duration Modulation.*

**Pulse-Position Modulation (PPM).** A form of pulse-time modulation in which the position in time of a pulse is varied.

**Pulse, Radio-Frequency.** See *Radio-Frequency Pulse.*

**Pulse Regeneration.** The process of restoring pulses to their original relative timings, forms, and/or magnitudes.

Note—In many devices, pulses may become distorted due to phase or amplitude distortion, limiting, or other processes. It is often desirable to restore the pulse to something resembling its original form before it has become so distorted that the original information which it contains is completely destroyed. This process is normally called pulse regeneration.

**Pulse Repetition Frequency.** The pulse repetition rate of a periodic pulse train.

**Pulse Repetition Period.** The reciprocal of the *Pulse Repetition Frequency.*

**Pulse Repetition Rate.** The average number of pulses per unit of time.

**Pulse Rise Time.** The interval between the instants at which the instantaneous amplitude first reaches specified lower and upper limits, namely, 10 per cent and 90 per cent of the peak-pulse amplitude unless otherwise stated.

**Pulse Spacing (Pulse Interval).** The interval between the corresponding pulse times of two consecutive pulses.

Note—The term "pulse interval" is deprecated because it may be taken to mean the duration of the pulse instead of the space or interval from one pulse to the next. Neither term means the space *between* pulses.

**Pulse Spectrum (Pulse Frequency Spectrum).** The frequency distribution of the sinusoidal components of the pulse in relative amplitude and in relative phase.

Note—The definition of this term was phrased to convey the idea that the spectrum is a complex (phasor) function of frequency and to express this function most nearly in a manner which corresponds to the method of measuring it (i.e., measuring amplitude and phase separately).

**Pulse Spike.** An unwanted pulse of relatively short duration superimposed on the main pulse.

Note—This term came into wide use in radar to define the first part of the pulse fed through a TR tube. This portion contains most of the pulse energy, has a duration about $10^{-3}$ that of the rest of the pulse, and an amplitude up to $10^6$ to $10^9$ times that of the rest of the pulse. Seen on a cathode-ray tube, it looks like a spike sticking up from the pulse. By extension, the term has come to be applied to any *unwanted* pulse of relatively short duration superimposed on the wanted pulse.

**Pulse Time, Leading Edge.** See *Leading Edge Pulse Time.*

**Pulse Time, Mean.** See *Mean Pulse Time.*

**Pulse-Time Modulation.** Modulation in which the time of occurrence of some characteristic of a pulse carrier is varied from the unmodulated value.

Note—This is a general term which includes several forms of modulation, such as pulse-duration, pulse-position, pulse-interval modulation.

**Pulse Time, Trailing Edge.** See *Trailing Edge Pulse Time.*

**Pulse Train.** A sequence of pulses.

**Pulse, Unidirectional.** See *Unidirectional Pulse.*

**Pulse-Width Modulation.** See *Pulse-Duration Modulation.*

**RMS (Effective) Pulse Amplitude.** The square root of the average of the square of the instantaneous amplitude taken over the pulse duration.

**Radio-Frequency Pulse.** A radio-frequency carrier amplitude-modulated by a pulse. The amplitude of the modulated carrier is zero before and after the pulse.

Note—Coherence of the carrier (with itself) is not implied.

**Trailing Edge Pulse Time.** The time at which the instantaneous amplitude last reaches a stated fraction of the peak pulse amplitude.

**Unidirectional Pulse.** A pulse in which pertinent departures from the normally constant value occur in one direction only.

Note—This is sometimes called "single-polarity" pulse, a term which is deprecated.

# Some Transistor Trigger Circuits*

PETER M. SCHULTHEISS†, ASSOCIATE MEMBER, IRE, AND HERBERT J. REICH†, FELLOW, IRE

**Summary**—Low *RC* time constants, small physical size, low operating voltages, and, in many units, a current gain in excess of unity make transistors promising for trigger-circuit applications.

Trigger action can be obtained with a single transistor if its current gain exceeds unity. Mathematical analysis indicates the possibility of triggering with successive pulses of the same polarity by use of an additional crystal diode. Improvement in some respects is obtained if a transistor is used in conjunction with a vacuum-tube cathode follower.

## INTRODUCTION

TRANSISTORS APPEAR very promising for use in trigger circuits for a variety of reasons. Their small input and output capacitances (1 to 2 $\mu\mu$f), in conjunction with low impedance levels, particularly in the emitter circuit, result in extremely small *RC* time constants and thus make possible fast rise times and high repetition rates. In practice, the inherent upper frequency limit of the transistor is more generally set by the hole transit time from emitter to collector than by these *RC* time constants. Small physical size, low operating voltages, and the absence of a stand-by power requirement are also attractive features. Selected transistors have a current gain exceeding unity, so that bistable operation can be obtained from a single unit. Most of the circuits discussed here make use of this property.

The practical usefulness of transistors is still limited severely by a number of factors. Production to close tolerances has thus far been impossible and even the performance of any one unit varies greatly over prolonged periods of time. Circuits operating well with one transistor may require drastic readjustments when another unit is substituted or when the same transistor is used after some lapse of time. Consequently, the present paper is to be considered as a survey of possibilities, rather than a development of circuits for immediate use in binary counters. Each of the circuits shown has been operated successfully with at least one transistor. Difficulties encountered in reproducing results with other transistors or altered parameters are noted in the discussion of individual circuits.

## BASIC CIRCUIT

The basic single-unit trigger circuit is shown in Fig. 1. The equivalent *T* network of Fig. 2 is constructed by use of conventional circuit theory and the standard definitions

$$\frac{\partial V_1}{\partial I_1} \equiv R_{11}, \frac{\partial V_1}{\partial I_2} \equiv R_{12}, \frac{\partial V_2}{\partial I_1} \equiv R_{21}, \text{ and } \frac{\partial V_2}{\partial I_2} \equiv R_{22}.$$

The characteristic determinant of this circuit is

$$\Delta = \begin{vmatrix} R_{11} + R_b + R_1 & R_{12} + R_b \\ R_{21} + R_b & R_{22} + R_b + R_2 \end{vmatrix}$$

and the condition for instability, and hence triggering, is therefore

$$(R_{11} + R_b + R_1)(R_{22} + R_b + R_2) < (R_{21} + R_b)(R_{12} + R_b). \quad (1)$$

As long as this inequality is satisfied anywhere within the possible operating range, trigger action can take place. It is seen immediately that both $R_1$ and $R_2$ tend to stabilize the circuit, so that triggering can be secured most easily when $R_1 = R_2 = 0$. This is a theoretical optimum since current ratings will, as a rule, not permit that mode of operation. If $R_1$ and $R_2$ are small and $R_b$ is permitted to become large, inequality (1) reduces to approximately

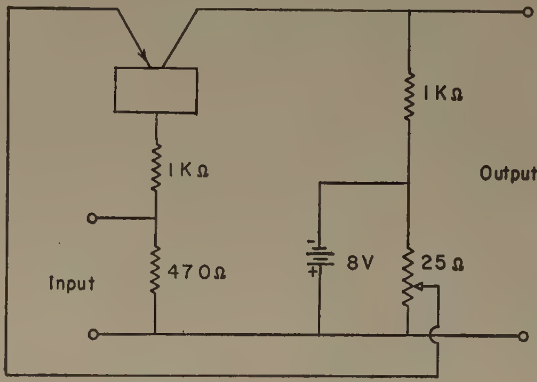

Fig. 1—Basic negative-resistance trigger circuit.

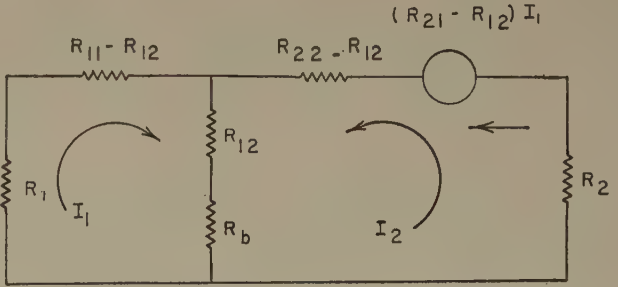

Fig. 2—Equivalent circuit for the circuit of Fig. 1.

$$\frac{R_{21}}{R_{22}} = \left.\frac{\partial I_2}{\partial I_1}\right|_{V_2 = \text{const}} > 1$$

since $R_{22}$ and $R_{21}$ are inherently large compared to $R_{11}$

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and $R_{12}$. A current gain in excess of unity is therefore seen to be a necessary condition for triggering.

An alternate point of view is to consider the negative resistance developed between base and ground. In order for triggering to be possible, the base circuit resistance $R_b$ must be larger in magnitude than the negative resistance. This point of view is particularly convenient because the negative-resistance characteristic can readily be viewed on an oscilloscope by means of the circuit of Fig. 3. As $R_1$ is raised, the negative resistance increases in magnitude and a larger value of $R_b$ is required. The minimum, however, is not raised appreciably in the absence of external resistance in the collector circuit, so triggering remains possible.

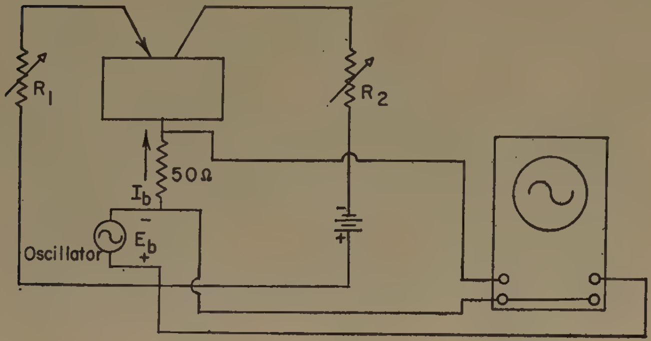

Fig. 3—Circuit for the oscillographic observation of negative-resistance characteristics.

Inequality (1) indicates that the introduction of resistance into the collector circuit should have little effect on circuit stability unless the resistance so introduced is comparable to $R_{22}$, which normally ranges from 20,000 to 60,000 ohms. Because of the nonlinearity of transistor characteristics, this prediction is not substantiated in practice. As current begins to flow through $R_2$ the effective collector-to-base voltage diminishes rapidly and the current gain falls below unity. The minimum of the negative-resistance characteristic is therefore raised (see Fig. 4), and often eliminated entirely, by collector-circuit resistances of the order of 1,000 ohms. This difficulty is, of course, most serious in low-voltage operation. For supply voltages under 10 volts the circuit of Fig. 5 was therefore found most successful.[1] It triggered reliably with 9 out of 10 transistors tried, without changes of circuit parameters except for bias adjustments. The circuit was tested in the frequency range from zero to 1 mc with a 6-volt collector supply. Output voltages of 2 volts with rise and decay times of the order of 0.1$\mu$sec were obtained. The trigger pulses can be introduced either at the emitter, as indicated in Fig. 5, or across the base-circuit resistance. A monostable version of this circuit is easily obtained by replacing the emitter-circuit resistance by a capacitor.

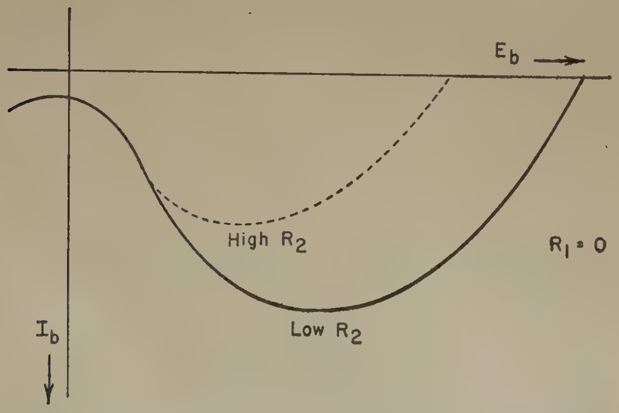

Fig. 4—Curves showing the effect of collector-circuit resistance.

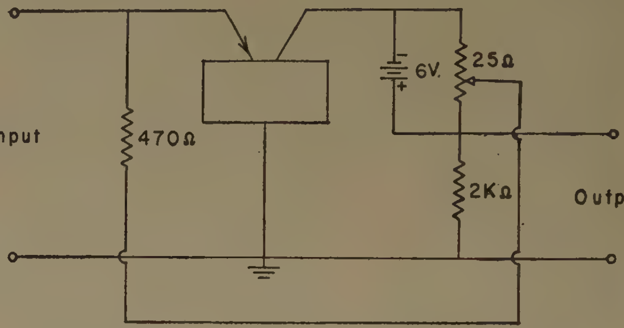

Fig. 5—Trigger circuit for use with low supply voltage.

## TRIGGER CIRCUITS OPERATING ON PULSES OF SINGLE POLARITY

The chief disadvantage of the circuit shown in Fig. 5 is the requirement of trigger pulses of alternating polarity. This difficulty can be overcome by means of a balanced circuit of the Eccles-Jordan type, such as that described by Eberhard, Endres, and Moore.[2] However, the requirement of a pair of transistors with at least reasonably similar characteristics makes the reproduction of results difficult with transistors chosen at random. Effort was therefore devoted to the development of single-input-polarity trigger circuits using only one transistor. One such circuit is included in the paper by Eberhard, Endres, and Moore. Its action depends upon the differentiation of the trailing edge of the input pulse by means of an $RC$ network, so that triggering is actually accomplished by means of pulses of alternating polarity. This scheme has the disadvantages of dependence, to some degree at least, on the wave shape of the input pulse and of an inherent frequency limitation well below the capabilities of the transistor alone (a figure of 250 kc is given in the paper). Work was therefore directed toward the development of circuits that do not incorporate energy-storing elements other than inherent parasitic capacitances.

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[1] Note that the 2,000-ohm resistor carries a current equal to the sum of collector and emitter currents, under the assumption of the polarities defined in Fig. 2. It therefore has the effect of a base resistance $R_b$.

[2] E. Eberhard, R. O. Endres, and R. P. Moore, "Counter circuits using transistors," *RCA Rev.*, vol. 10, pp. 459–476; December, 1949.

In all the circuits tried, the attempt was made to introduce positive pulses alternately at the emitter and the base. This commutation of pulses was accomplished by one or two diodes biased by means of voltages from the proper points in the circuit. One circuit of this type, which was operated successfully, is shown in Fig. 6. It produced a square-wave output of 30 volts peak-to-peak amplitude in the frequency range up to 600 kc (400 kc for reasonably good wave shape). The relatively low upper frequency limit is probably due largely to the wide range of collector current variation (indicated by the large output amplitude) that is necessary to produce a base-voltage swing large enough to cut off the diode. Experimental evidence indicates that the rise time increases with the magnitude of the output voltage.

The circuit of Fig. 6 was designed as follows: The negative-resistance characteristic (base voltage versus base current) was observed by an oscilloscope by means of the arrangement shown in Fig. 3. Circuit parameters were then adjusted until the trace exhibited a long linear section of negative resistance. Mathematical considerations that will be presented indicate that the linearity of the characteristic is essential to proper performance. Characteristics of this type are obtained most readily from transistors with high current gain by the introduction of resistance into the emitter circuit. Once the desired shape had been obtained (Fig. 7) the oscilloscope was calibrated and the base-circuit resistance chosen somewhat larger than the reciprocal of the slope of the linear portion of the negative-resistance characteristic. The base-circuit bias was then adjusted to bring the diode to the point of conduction at the upper stable point. Calibration of negative-resistance characteristics in the manner described could easily be made accurate to within 10 per cent.

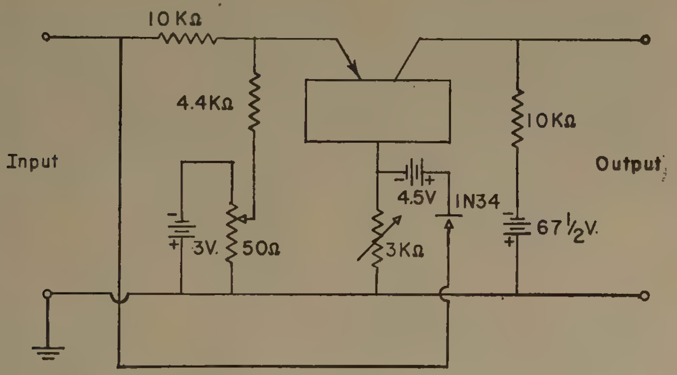

Fig. 6—Trigger circuit operating on single-polarity pulses.

The circuit proved not at all critical with respect to adjustment of emitter- and base-circuit resistance when a transistor with the proper characteristics was used.

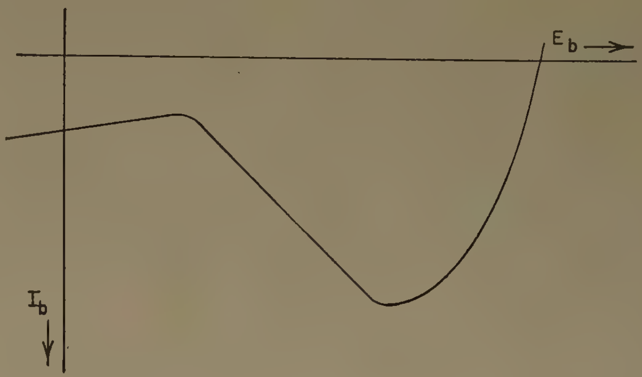

Fig. 7—Desired negative-resistance characteristics.

Unfortunately, only one transistor with sufficiently linear negative resistance was available and attempts to reproduce the results with other transistors proved unsuccessful. It was hoped that better linearity could be secured by eliminating the collector resistance entirely and increasing the emitter and base resistances as shown in Fig. 8. In this circuit the output voltage must clearly be taken across the base resistance and the larger voltage swing at that point greatly simplifies the problem of biasing the diode to cutoff. The upper (high-current) stable point would probably lie far beyond the ratings of the transistor if it were located at the intersection of the negative-resistance characteristic with the load line corresponding to $R_b$ ($MN$ in Fig. 10). Fortunately, the effective load line ($M'N'$ in Fig. 9) is determined by the parallel combination of $R_b$ and the resistance in series with the diode as soon as the latter begins to conduct. This load line determines the upper stable point, which can therefore be adjusted by controlling the bias voltage in series with the diode.

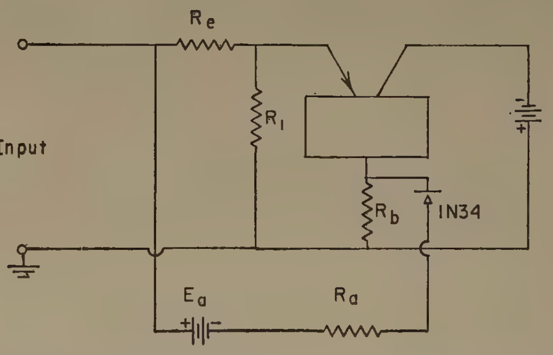

Fig. 8—Modified circuit for operation with single-polarity pulses.

## ANALYSIS OF CIRCUIT

The following paragraphs give a summary of mathematical considerations for the determination of optimum circuit parameters. Linearity of the negative-resistance characteristic is assumed in the neighborhood of the unstable equilibrium point and the output impedance of the pulse source is assumed to be zero. The analysis is applicable not only to the circuit of Fig. 8 but also that of Fig. 6, except that for the circuit of Fig. 6 the upper stable point will be determined by the intersection of the negative-resistance line with the load line $MN$ corresponding to $R_b$ and, as a result, the pulse amplitude required for triggering at the base may be somewhat larger than computed here, particularly if the diode is still biased beyond cutoff at the upper stable point.

The nomenclature of circuit parameters is defined in Fig. 9, in which $1/R_n$ is the slope of the linear portion of the negative-resistance characteristic. Assume that the circuit is initially in the low-conduction stable state. When a positive pulse of magnitude $E_1$ is applied to the input, the negative-resistance characteristic is shifted to the left by an amount $E_1R_1/(R_1+R_e)$.[3] If this shift carries the negative-resistance curve past the position of tangency with the load line $R_b$, the circuit will trigger. No pulse can reach the base during this process since the diode is cut off.

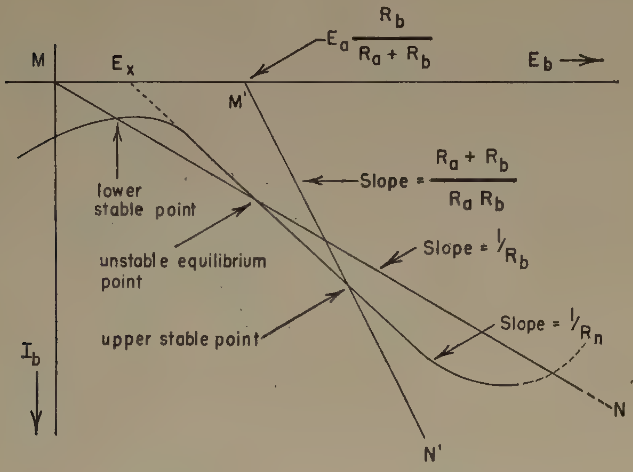

Fig. 9—Negative-resistance characteristic and load lines.

Assume that the circuit has been triggered as described above. The diode begins to conduct when the instantaneous base voltage assumes the value $E_a$. At that instant the load line changes to one of slope

$$\frac{R_a+R_b}{R_aR_b}$$

with a terminus on the horizontal axis at the point where $E_b=E_aR_b/(R_a+R_b)$. Hence the high-conduction stable point may be determined from the equation of the negative-resistance line

$$E_b = R_nI_b + E_x \tag{2}$$

and that of the load line $M'N'$

$$E_b = \frac{R_aR_b}{R_a+R_b}I_b + E_a\frac{R_b}{R_a+R_b}. \tag{3}$$

Solution of these two equations yields the following relation for the base voltage at the upper (high-conduction) stable point:

$$E_b = \frac{E_a - \frac{R_a}{R_n}E_x}{1-\frac{R_a}{R_n}\left(1-\frac{R_n}{R_b}\right)}. \tag{4}$$

The base voltage at the upper stable point can evidently not be smaller than $E_a$, since the diode will not begin to conduct before this base voltage is reached and no second stable state can therefore be reached under the assumption of linearity. The base voltage corresponding to the upper stable point is equal to $E_a$ either in the trivial case when $R_a=0$ or else when

$$E_a = \frac{E_x}{1-\frac{R_n}{R_b}}. \tag{5}$$

This last expression may be shown to give also the base voltage at the unstable equilibrium point. Hence an infinitesimal voltage would be sufficient to retrigger the circuit if $E_a$ were adjusted to that value. This prediction is substantiated by the results of the more general analysis.

Assume that the system has come to rest at the upper stable point given in general by (4). In order to cause retriggering, the pulse applied to the base must be large enough (a) to cause the base voltage to take on a value smaller than its magnitude at the unstable equilibrium point and (b) to cut off the diode in the base circuit. The pulse simultaneously applied to the emitter causes the negative-resistance characteristic to shift to the left. This fact is included in the calculations by replacing $E_x$ by the term $E_x-E_iR_1/(R_1+R_e)$, in which $E_i$ is the magnitude of the applied pulse.

(a) The pulse required to reach the unstable equilibrium point is given by the solution of the equation

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[3] The collector supply voltage is assumed to be large compared to the change in base voltage caused by the applied pulse. If this is true, the shift of the negative-resistance characteristic is determined by the effective emitter-circuit voltage, as computed by Thevenin's theorem.

$$\frac{E_x - \frac{R_1}{R_1 + R_e} E_i}{1 - \frac{R_n}{R_b}} = \frac{(E_a - E_i) - \frac{R_a}{R_n}\left(E_x - \frac{R_1}{R_1 + R_e} E_i\right)}{1 + R_a\left(\frac{1}{R_b} - \frac{1}{R_n}\right)}. \qquad (6)$$

The left side of (6) gives the value of $E_b$ at the unstable equilibrium point; the right side gives the value of $E_b$ at the upper stable point, i.e., the instantaneous operating point with a voltage $E_i$ applied to the input. Solution of (6) gives the following expression for $E_i$:

$$E_i = \frac{E_a\left(1 - \frac{R_n}{R_b}\right) - E_x}{1 - \frac{R_n}{R_b} - \frac{R_1}{R_1 + R_e}}. \qquad (7)$$

(b) The pulse required to cut off the diode is given by the solution of the equation

$$\frac{(E_a - E_i) - \frac{R_a}{R_n}\left(E_x - \frac{R_1}{R_1 + R_e} E_i\right)}{1 + R_a\left(\frac{1}{R_b} - \frac{1}{R_n}\right)} = E_a - E_i.$$

The left side of this equation represents the instantaneous value of $E_b$. The diode must cut off when the (negative) voltage $E_b$ on its cathode becomes equal to the (negative) voltage $E_a - E_i$ on its anode. Solution of the above equation gives the following expression for $E_i$:

$$E_i = \frac{E_a\left(1 - \frac{R_n}{R_b}\right) - E_x}{1 - \frac{R_n}{R_b} - \frac{R_1}{R_1 + R_e}}. \qquad (8)$$

Equations (7) and (8) are clearly identical. This fact indicates that the diode will always cut off when the system reaches the unstable equilibrium point. The required pulse magnitude is seen to be independent of $R_a$. Since a large resistance $R_e$ must be used in series with the emitter in any case, the operation of the circuit should be fairly independent of the impedance of the pulse source. This prediction is borne out by experimental results. If $E_a = E_x/(1 - R_n/R_b)$ the required pulse amplitude approaches zero in the limit.

Equations (7) and (8) indicate some of the difficulties in the construction of a circuit of this type. In general, $E_a \geqq E_x/(1 - R_n/R_b)$ so that the numerators are inherently positive and the following condition is necessary for triggering with a positive pulse:

$$\frac{R_n}{R_b} + \frac{R_1}{R_1 + R_e} < 1. \qquad (9)$$

Hence the design problem reduces essentially to a proper assignment of the relative magnitudes of $R_n/R_b$ and $R_1/(R_1 + R_e)$. In general, it is desirable to make $R_b$ nearly equal to $R_n$. The ratio $R_1/(R_1 + R_e)$ must then be small, however, so a very large pulse is required for triggering at the emitter. On the other hand, if $R_1/(R_1 + R_e)$ is made relatively large, the unstable equilibrium point cannot readily be located at a reasonably high base voltage without requiring excessive pulse amplitudes for triggering at the emitter. This difficulty is accentuated greatly by any nonlinearity of the negative-resistance characteristic.

Experimentally, circuits constructed on the basis of these considerations showed a considerable tendency to retrigger twice, as shown by Fig. 10. This phenomenon is not entirely unexpected, inasmuch as a small pulse at the base is sufficient to trigger the circuit. Unless the trigger pulse is very fast compared to the system response, the pulse applied simultaneously to the emitter will cause retriggering.

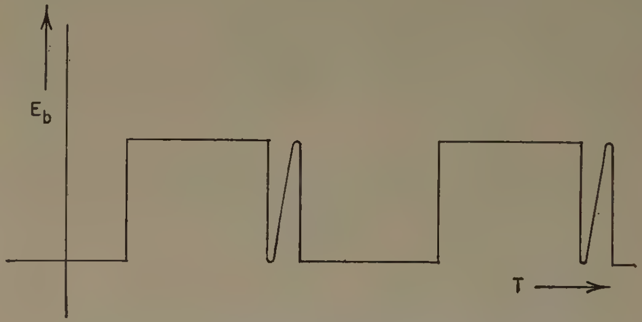

Fig. 10—Typical base-voltage waveform for the circuit of Fig. 8.

## Other Circuits

The voltage gain of the type-A transistor far exceeds its current gain, but in view of the low emitter-circuit impedance, an effective gain greatly in excess of the current gain cannot be readily realized when the collector

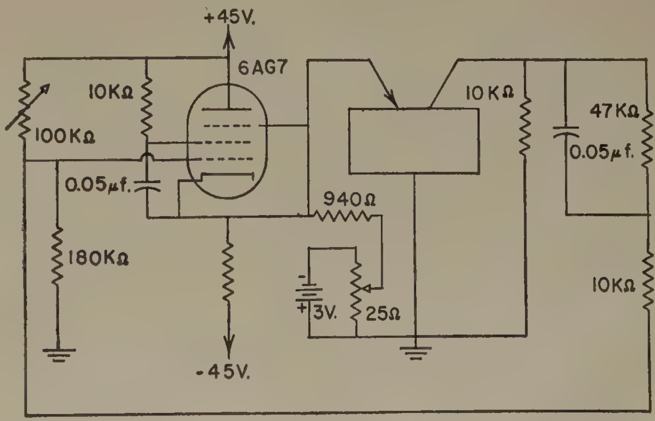

Fig. 11—Trigger circuit using a cathode-follower as an impedance-matching element.

is coupled to the emitter circuit. A vacuum-tube cathode-follower amplifier was used in an effort to overcome this difficulty, the resulting circuit taking the form shown in Fig. 11. The introduction of vacuum tubes has, of course, some inherent disadvantages, but since a cathode-follower with good frequency response up to at least 10 mc is easily constructed, some improvement in response speed over that of the simple transistor circuit can be expected from the combination. The circuit of Fig. 11 consists essentially of a standard transistor amplifier, the cathode follower providing the necessary impedance match between collector and emitter circuits. The introduction of the coupling network between collector and cathode-follower grid was necessary in order to reduce the dc loop gain. Without this modification, an excessively large trigger pulse (in the order of 5 volts) was required. At high frequencies, the transfer ratio of this network is essentially unity, so that none of the advantages of the circuit have been sacrificed.

The circuit of Fig. 11 triggered on pulses of alternating polarity applied to the emitter at repetition frequencies up to at least 2 mc (the input pulse at these frequencies was very nearly a sine wave, so that the results are not easily interpreted). A trigger pulse of less than 0.5 volt was found adequate. The output amplitude was between 3.5 volts and 6.5 volts, depending upon operating conditions. The circuit appears promising, as its performance is not critically dependent upon transistor characteristics. The results were readily reproduced with a second transistor with only slight bias readjustments. Loop-gain measurements indicated gain values above unity up to frequencies somewhat in excess of 10 mc without any attempt at compensation. The achievement of repetition rates of that order of magnitude therefore appears a distinct possibility.

## LIMITATIONS

Reference has already been made to the difficulty of reproducing certain experimental results with other transistors or at a later time. Circuits depending entirely upon the value and linearity of the current gain are most troublesome from that point of view. Negative-resistance characteristics vary widely from unit to unit and tend to change as a function of time for any given unit. Behavior in that respect is quite unpredictable.[4] A transistor may maintain its characteristics closely under intermittent use for a period of many weeks and then suddenly, without overload, deteriorate to the point of uselessness in the space of a few hours. In some instances deterioration on the shelf was more rapid than when a transistor was used regularly in circuits. Some bias readjustment is generally necessary even to restore to operation a trigger circuit that has been turned off only a few minutes previously. As a rule, however, once trigger action commences, no further adjustment is necessary to assure continued operation. All of these difficulties were minimized in circuits using the voltage gain, rather than the current gain, of the transistor, because the impedance difference between emitter and collector circuits maintains the voltage gain at a reasonably high level even after the current gain has fallen below unity.

Some transistors whose current gain has fallen off can be reactivated by applying between collector and the base a positive voltage pulse from a charged capacitor. Results from such treatment are, however, by no means uniformly positive. In some instances the negative-resistance characteristic showed a tendency to develop two distinct minima, as shown in Fig. 12. By means of careful pulsing, this phenomenon could be accentuated considerably, so that the construction of a trigger circuit with three stable states became a possibility. One such circuit was constructed and found to trigger as anticipated. The lack of a reliable method for obtaining such characteristics and their long-term instability have so far discouraged further work in that direction.

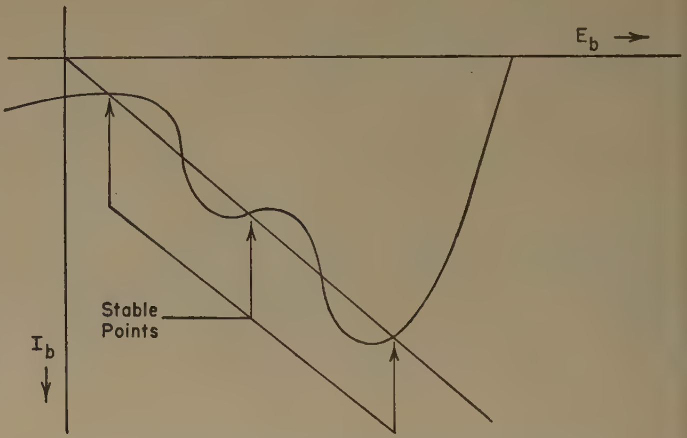

Fig. 12—General form of negative-resistance characteristic used in a tristable trigger circuit.

## ACKNOWLEDGMENT

The work which forms the basis of this paper was done under the sponsorship of the Office of Naval Research. The authors also wish to acknowledge the assistance of H. Epstein and C. Susskind in certain portions of the work.

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3. H. J. Reich and R. Ungvary, "A transistor trigger circuit," *Rev. Sci. Instr.*, vol. 20, pp. 586–588; August, 1949.

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[4] Improvements in the design and manufacture of transistors may have greatly reduced this difficulty during the time that has intervened between the experimental work and the publication of this paper.

# Condenser Microphone Sensitivity Measurement by Reactance Tube Null Method*

H. E. VON GIERKE† AND W. W. VON WITTERN†

*Summary*—**A reactance tube circuit acting as a sinusoidally varying capacity in the frequency range from 20 to 200,000 cps is used to compensate the capacity variations of a condenser microphone in a high-frequency carrier circuit. The microphone is driven as in the electrostatic actuator method by a voltage of the same frequency, that modulates the reactance tube capacity. By adjusting amplitude and phase of the reactance tube modulation, the total capacity change of the circuit becomes zero. The sensitivity and phase lag of the microphone response can be determined from the actuator and modulating voltages when the circuit is properly calibrated. The circuit is advantageous in obtaining an equivalent pressure chamber calibration of condenser microphones up to 200 kc. The null method makes it useful for relatively unsensitive microphones.**

## INTRODUCTION

A CALIBRATED reactance tube circuit (RTC), designed to represent a variable condenser, is useful for the measurement of periodic capacity changes as in the calibration of condenser microphones.

Fig. 1 will be used to explain the principle of this calibration method. The microphone is driven as in the electrostatic actuator method.[1,2] The sinusoidal actuator voltage and the necessary polarizing dc voltage is applied between the membrane and the counter-electrode of the microphone, to produce periodic electrostatic forces which cause vibrations of the membrane producing capacity variations. The capacity of the microphone $C$ and the inductance of the coil $L$ form a circuit which is excited by a carrier frequency generator the frequency of which is situated on the side of the resonance curve of the circuit. A change of the capacity of the microphone changes the tuning of the circuit and thus the carrier frequency voltage across the circuit. This voltage is modulated by these capacity changes. When rectified and displayed on the oscilloscope, the curve represents the capacity change of the circuit. However, owing to the response of the $LC$ circuit, the amplitude of this trace depends on the modulating frequency.

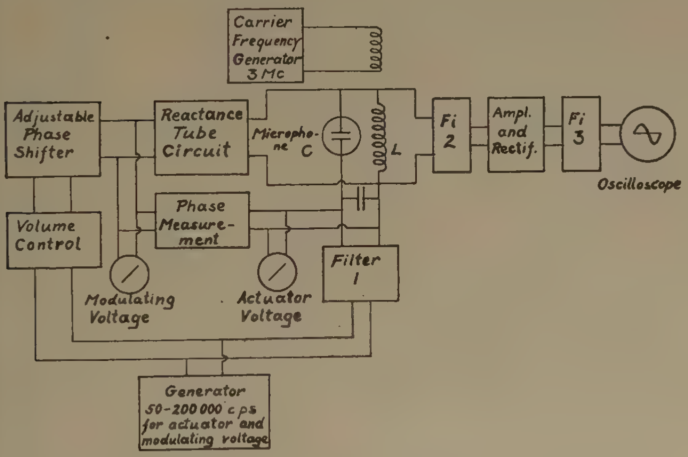

Fig. 1—Schematic diagram of the arrangement used for the microphone calibration by means of the reactance tube circuit.

Parallel to the $LC$ circuit is the reactance-tube circuit (henceforth referred to as RTC). Its modulating voltage is taken from the same generator as is the actuator voltage for the microphone. By adjusting the amplitude and phase of this modulating voltage the capacity change of the RTC can be made equal and opposite in phase to the capacity change of the microphone. Under the condition that any series resistances in the microphone and RTC branch are negligible the total capacity change of the circuit then will be zero. This adjustment can be controlled by observing the oscilloscope. The ratio of the modulating voltage to the actuator voltage is proportional to the sensitivity of the microphone. The phase relation between these voltages corresponds to the phase relation between membrane-force and resultant capacity change. This phase relation is easy to measure because the magnitude of these voltages is of the order of several volts.

The filter networks 1 and 2 prevent the actuator voltage from influencing the oscilloscope directly by electrical cross talk. The output impedance of the filter 1 must be very low for the carrier frequency. Filter 3, sharply tuned to the frequency under measurement, is applied to decrease the noise level so that a very high amplification can be used.

The advantage of this null method is that the result is independent of the frequency response characteristics of filters and amplifiers and of the frequency response characteristic which results from transients of the $LC$ circuit.

## THE REACTANCE TUBE CIRCUIT

For such measuring purposes the RTC must produce only reactance changes without resistance changes because both produce a carrier modulation in the circuit. A constant ohmic resistance parallel to the circuit would have no influence other than on the sensitivity of the measurement.

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* Decimal classification: R254.2×R385.53. Original manuscript received by the Institute, May 8, 1950; revised manuscript received, October 30, 1950.

A preliminary report of this work was presented before the Acoustical Society of America in St. Louis, November 19, 1949, JASA 22 (1950) p. 85.

† Aero Medical Laboratory Engineering Division, Air Materiel Command, Dayton, Ohio.

[1] M. Gruetzmacher and E. Meyer, "Eine Schallregistriervorrichtung zur Aufnahme der Frequenz-Kurven von Telephonen und Lautsprechern," *Elekt. Nach. Techn.*, vol. 4, pp. 203–211; 1927.

[2] L. J. Sivian, "Absolute calibration of condenser transmitters," *Bell Sys. Tech. Jour.*, vol. 10, pp. 96–115; January, 1931.

A pure reactance exists between plate and cathode of an electron tube when the plate voltage is "fed back" to the grid with a 90° phase shift.[3,4] To realize this condition in the RTC (Fig. 2) we employ an adjustable phase shifting stage. To produce a periodically changing capacity the amplitude of the reactance tube (RT) plate current is modulated. Direct-grid modulation of the RT produces a varying plate resistance in parallel with the desired varying capacity. The modulation is therefore produced in a special stage. To avoid undesired variations of the plate resistance of the RT, the modulating voltage is removed by compensation and filtering following the "modulation stage."

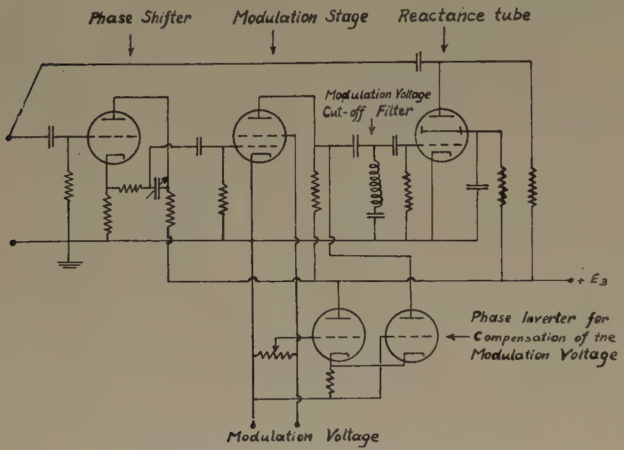
Phase Shifter
Modulation Stage
Reactance tube
Modulation Voltage Cut-off Filter
+ $E_3$
Phase Inverter for Compensation of the Modulation Voltage
Modulation Voltage

Fig. 2—Complete circuit of the reactance tube circuit without resistive component.

### PHASE ADJUSTMENT OF THE RTC

If in addition to the varying capacity, the RTC has a varying resistance component the carrier voltage across the tuned circuit *LC* (Fig. 1) is modulated when the carrier frequency is located on the side of the resonance curve, and also when the same carrier frequency is located at the peak of the resonance curve. But a small change in capacity alone will modulate the voltage only when the carrier frequency is located on the side of the resonance curve. Therefore, by switching off the microphone actuator voltage and tuning the *LC* circuit exactly to the carrier frequency, the phase of the RTC can be adjusted so that the modulation by the RTC disappears. This indicates that the circuit is acting as a pure reactance. Exact tuning of the circuit is obtained by adjusting for disappearance of the modulation produced by the microphone, when the modulating voltage of the RTC is cut off.

### MICROPHONE CALIBRATION PROCEDURE

Following the adjustment of the RTC, for calibration of the microphone, the *LC* circuit is slightly detuned with the carrier frequency maintained constant. The actuator voltage is now applied resulting in modulation of the voltage across the *LC* circuit by both the microphone and the RTC. By adjusting the phase and the amplitude of the modulating voltage to the RTC, the summed carrier modulation is reduced to zero. In this compensated state no sideband frequencies are present at the input of the RTC. This condition is essential because sideband frequencies would introduce errors, since the RTC is not adjusted to them. Now only the unmodulated carrier to which the RTC acts as a pure capacity is present.

### FREQUENCY RESPONSE CHARACTERISTIC OF THE RTC

The capacity change of the RTC must be the same for all modulating frequencies. This means that the modulation percentage of the output current resulting from a certain modulating voltage must be independent of the modulating frequency. The frequency response characteristic of the circuit, between the modulating stage and the output stage, is therefore designed to be flat between the sideband frequencies of the modulated carrier. This is accomplished by resistance capacity coupling even for the 2- to 4-mc range because the tubes do not have to produce any amplification. Therefore very low coupling resistances ($\sim$200 ohms) can be used. The modulation percentage was checked by measuring the carrier current peaks of the output stage with and without a modulating voltage. Its frequency response characteristic was found to be flat within less than $\pm 0.5$ db in the frequency range 50–200,000 cps.

### CAPACITY CALIBRATION OF THE RTC

The capacity change of the RTC resulting from the application of a certain modulating voltage was calibrated at a low frequency by replacing the microphone with a variable plate condenser driven by an electrodynamic moving coil system. This system was driven by the generator producing the modulating voltage of the RTC. The condenser plate amplitude was controlled by a microscope. The extreme values of its capacity due to this amplitude were measured under static conditions. Because the capacity changes obtained this way were much larger than the capacity changes of the microphone, they were reduced by means of a condenser network.

### DETERMINATION OF THE PRESSURE EQUIVALENT TO THE ACTUATOR VOLTAGE

The microphone calibration described gives the capacity change of the microphone $\Delta C_M$ due to the actuator voltage. To obtain the capacity change resulting from a sound pressure $P$ it is necessary to determine, at one frequency, what sound pressure is equivalent to a certain actuator voltage. In the calibration circuit described above (Fig. 1) with the RTC switched off, the microphone is simultaneously excited by an actuator

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[3] C. Travis, "Automatic frequency control," PROC. I.R.E., vol. 23, pp. 1125–1142; October, 1935.

[4] P. Guettinger, "Frequenzmodulation," Zuerich, Leeman, p. 111; 1947.

voltage and by a known low-frequency sound pressure produced in a pressure chamber. The piston of the pressure chamber is driven by an electrodynamic speaker system excited by the actuator voltage generator. By adjusting the amplitude and the phase of the current in the speaker system, the capacity change of the microphone can be made zero. The forces resulting from the actuator voltage and from the sound pressure are equal and opposite in phase in this case.

## RESULTS OF THE MICROPHONE CALIBRATION

With the factor $\Delta C_M/P$ determined, the open circuit voltage $E_M$ resulting from the pressure $P$ in the usual low-frequency circuit with the battery voltage $E_B$ can be calculated when the static capacity $C$ of the microphone is measured:

$$\frac{E_M}{P} = E_B \frac{\Delta C_M}{P} \frac{1}{C}.$$

Fig. 3 shows the results of the calibration of a 640 AA microphone and a small condenser microphone of our own design.[5] To show sensitivity and phase they are presented as polar diagrams. The radial scale indicates the sensitivity of the microphone in decibels, the angles of the radii indicate the phase lag between the pressure acting on the membrane and the output voltage of the microphone. The frequencies are distributed along the curve in a way which depends on the qualities of the microphone. Every loop indicates a resonance in this kind of diagram. The calibration method described represents a pressure calibration. Its results correspond within 1 db with pressure chamber calibrations obtained with the usual methods, however not extended to frequencies that high. They also correspond to free field reciprocity calibrations within 1 db in the low-frequency range, and show the deviation to be expected at high frequencies due to the diffraction on the microphone.

## APPLICATION OF THE CALIBRATION METHOD

The calibration method described might be especially useful for the calibration of relatively unsensitive condenser microphones designed for the measurement of high pressures and shock waves. For this calibration it is hardly possible to produce by conventional methods sinusoidal pressures of the necessary magnitude and frequency. By the electrostatic actuator method too only limited forces can be produced on the membrane, however not limited in frequency. Even with these limited forces a calibration of insensitive microphones will be possible, because the method described allows diminution of the noise level and a considerable increase in sensitivity by the application of very sharp filters without increasing the difficulties of the measurement.

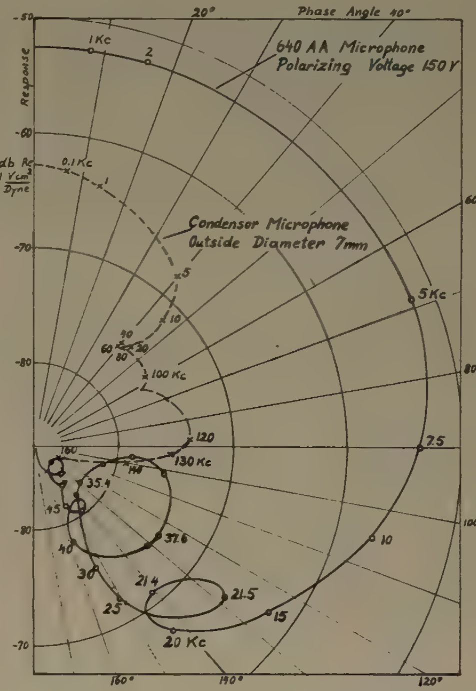

Fig. 3—Sensitivity and phase lag of two different microphones:
radial scale = sensitivity in db ref. 1 volt cm²/dyne
angle of the radii = phase lag in degrees between pressure acting on the membrane
output voltage numbers along the curves = frequencies in cps.

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[5] H. F. Koster, H. E. von Gierke, and H. L. Oestreicher, "The calibration of microphones to one hundred kilocycles per second," *Jour. Acous. Soc. Amer.*, vol. 21, p. 58; January, 1949.

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# Grid Current and Grid Emission Studies in Thyratrons—the Trigger-Grid Thyratron*

L. MALTER†, SENIOR MEMBER, IRE, AND M. R. BOYD‡, STUDENT, IRE

*Summary*—**Grid emission leads to unstable control characteristics in thyratrons. Positive ion grid current during the afterglow results in delayed recovery of grid control. Both these drawbacks can be overcome by use of low values of grid resistors, but at the expense of input signal current sensitivity. The requirement for low grid resistance can be eliminated by assigning the firing function to a fine wire electrode known as the trigger grid. The other function of a conventional thyratron grid, which is to promote recovery of control, is assigned to another electrode known as the blocking grid. Since the latter electrode is connected to a bias supply through a low resistance, its emission and positive ion collection do not produce disturbing effects. Comparisons of an experimental trigger grid thyratron with its conventional prototype reveal many advantages of the former.**

## INTRODUCTION

A THYRATRON desideratum is the reduction of the time needed for recovery of grid control following the interruption of a discharge. This time, which is generally referred to as "deionization time," we choose to call "recovery time," $T_R$, in view of the fact that recovery of grid control can occur when intense ionization is still present in the tube.[1] The period following the interruption of a discharge is generally referred to as the "afterglow."

One of the factors that delays recovery of grid control is the flow of positive-ion current to the grid during the afterglow. This current produces a potential drop in the grid resistor which neutralizes the effect of the grid bias source, thus preventing the early return of the grid potential to its bias value. This can be minimized by a reduction in the effective area of the grid, thus reducing its positive-ion collection. It is the purpose of this paper to describe work in progress along these lines. As will be seen, the step of reducing grid area is effective not only in reducing recovery time, but is accompanied by additional benefits, such as (1) increased sensitivity to input signal current, (2) reduced effects of grid emission with consequent increased stability as regards variation in heater current and plate power, and (3) reduced interelectrode capacitances.

## CONSIDERATIONS REGARDING THYRATRON RECOVERY

### A. *Plasma Decay*

Fundamental studies of phenomena occurring during the afterglow[1] have led to interesting results that are of value in understanding thyratron behavior. Significant among these results are a conception of a process of plasma decay and the conditions necessary for recovery. The plasma decay has been shown to be due primarily to a diffusion process.

A gas plasma is a region of essentially equal positive ion and electron density. Fundamental studies indicate that when the discharge in a thyratron is interrupted the tube is substantially completely suffused with a plasma, and that as the ionization then diminishes it does so in such a manner that the plasma retains its plasma-like characteristics. This is accomplished by an equal rate of loss of electrons and ions to the walls and electrodes.

The plasma decay can be speeded up by decreasing the molecular weight and pressure of the gas and by shrinking down the geometry of the structure. In general, rapid recovery of grid control requires a rapid decrease of plasma density. Decreasing the pressure or using a lighter gas may lead to difficulties resulting from more rapid cleanup of the gas present, thus resulting in shorter life. Shrinking the geometry can take the form of decreasing interelectrode spacing. This speeds up the diffusion process whereby deionization takes place.

The above described factors which determine $T_R$ may be described as primary since they are inherently properties of the tube geometry or its gas filling, and are effective through the mechanism of altering the rate of plasma decay. $T_R$ can also be varied by changes in operating parameters, such as anode current or voltage, grid voltage, and grid resistance. These are secondary factors which do not materially influence the rate of plasma decay. It will be of interest to examine the role of the grid in promoting recovery. This will throw light on the manner in which recovery is also dependent upon the other operating parameters.

### B. *The Role of the Grid in Recovery*

The grid in a conventional thyratron triode plays a double role. One function is to determine whether or not a discharge is to occur in a previously unfired tube. Thus, it is a firing electrode. An equally important function is to perform in such a fashion as to promote or insure recovery of control, following the interruption of a discharge. This matter, which is fully discussed in the literature[1], is briefly presented here.

As was stated above, during the afterglow the tube is suffused with a plasma. Now if the grid is negative with respect to the plasma, it will surround itself with a positive-ion sheath. As the plasma decays, the sheath thickness increases. The positive-ion current to the grid is determined by the rate at which ions diffuse out of the

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† RCA Laboratories, Princeton, N. J.
‡ Stanford University, Stanford, Calif.

[1] L. Malter and E. O. Johnson, "Studies of thyratron behavior," *RCA Rev.*, vol. 11, p. 165; June, 1950.

plasma into the sheath. Since the sheath area surrounding the grid varies only slightly with grid potential, one finds the positive-ion current to the grid to be almost independent of grid potential.[2] This in turn implies that the density of ionization in the plasma will be practically uninfluenced by the grid potential.

As the plasma decays, the sheath around the negative grid increases in thickness until it finally extends across the grid opening or openings.[3] When this occurs, the residual plasma in the anode and cathode spaces becomes independent and can assume widely different potentials. Each plasma then assumes a potential very close to that of the most positive electrode with which it can make contact.[3] Thus, if the grid is negative with respect to the cathode and anode, the plasma potential on the cathode side will be substantially that of the cathode and similarly on the anode side the plasma potential will be close to that of the anode. The potential distribution along a section through the cathode-grid aperture is as shown in Fig. 1. The tube now behaves like a thyratron with reduced spacings, the "effective cathode" being at *A* and the "effective anode" at *B*. Whether or not the tube fires depends closely on whether or not a thyratron with cathode at *A*, anode at *B*, and grid at *G*, and with the potentials as shown in Fig. 1 will refire. As time progresses, the grid sheath expands, forcing *A* and *B* farther away from the grid. Thus, the effective refiring geometry varies in the fashion that one would expect, to explain the observed results that $T_R$ increases with anode refiring voltage. It is also apparent that since the residual plasma density at the discharge interruption is a monotonic function of the discharge current $T_R$ should increase with anode current. As the grid sheath expands, the control ratio gets larger, approaching as a limit the control ratio of the unfired tube.

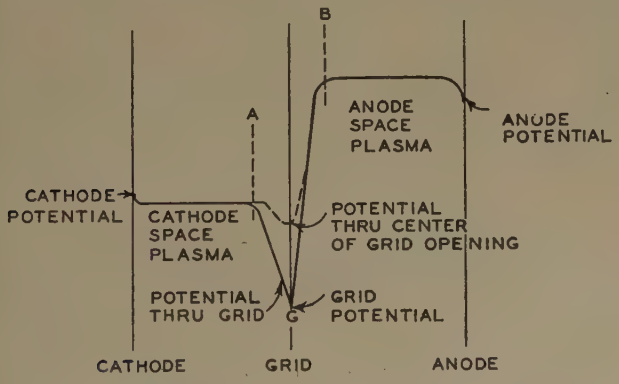

Fig. 1—Potential distribution between thyratron electrodes.

For the grid to recover control, it must first surround itself with a sheath of sufficient extent to block its own openings. The more negative the grid, the more rapidly does the sheath block the openings and the greater the spacings from grid to the effective anode and cathode positions. This explains why recovery time is reduced by making the grid increasingly negative.

In the conventional triode thyratron the dual functions of firing and of promoting recovery are performed by the same electrode, the control grid. By assigning these functions to separate electrodes one can achieve improved performance in several ways.

Note—There is a common misconception that the decrease of $T_R$ brought about by making the grid increasingly negative arises from a more rapid deionization of the plasma. Evidence for the view that the grid potential has no first-order effects on the rate of deionization is presented in the literature.[1]

### C. Grid Resistance and Recovery

We are now in a position to see how the presence of a grid resistor causes an increase in $T_R$. (We summarize here briefly some of the experimental results and conclusions of the literature.[1])

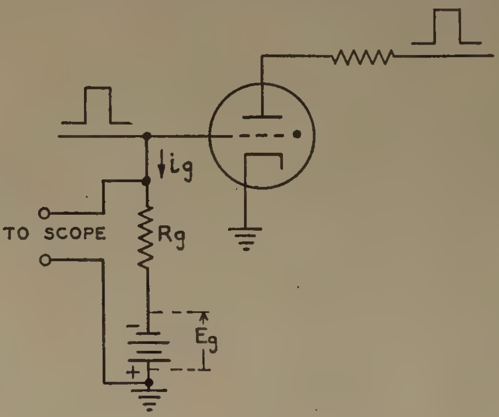

Fig. 2—Thyratron study test circuit.

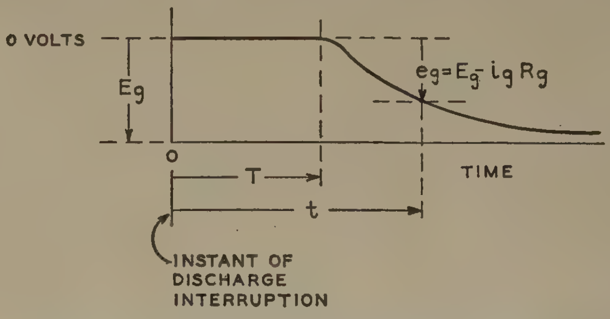

Fig. 3—Thyratron grid voltage during period following interruption of discharge.

A three-element thyratron is first fired by simultaneous pulsing of plate and grid. (See Fig. 2.) The grid potential during the afterglow is found to follow a course as shown in Fig. 3. Following the interruption of the discharge, the grid remains substantially at cathode potential for time *T*, and then decays along an exponential curve towards the bias value. This is explained in the following manner: Following the discharge interruption, positive-ion current flows to the grid, this current being given by

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[2] I. Langmuir and H. M. Mott-Smith, "Studies of electric discharges in gases at low pressures," *GE Rev.*, vol. 27, pp. 449–455, 538–548, 616–623, 762–771; 1924.

[3] W. Koch, "Experimental demonstration of the existence of ion sheaths at electrodes of rectifiers during deionization," *Zeit. fur Tech. Phys.*, vol. 17, p. 446; 1936.

$$i_g = i_{g0}e^{-t/\tau}, \tag{1}$$

where $i_{g0}$ is positive-ion current to probe immediately following the interruption of the discharge, and $\tau$ is the time constant of the plasma decay.

When

$$i_gR_g > E_g \tag{2}$$

(these quantities are defined in Figs. 2 and 3), the drop in the grid resistor exceeds the bias value and the grid tends to go positive. Under this condition, the grid also collects electrons, the electron current being just sufficient to make the grid operate close to cathode potential. This condition persists until

$$i_gR_g = E_g. \tag{3}$$

Beyond this point ($t = T$, in Fig. 3) the grid stops collecting electrons and becomes increasingly negative.

There is no possibility for the recovery of grid control until $t > T$. This follows from the fact that when $t < T$ the anode and cathode plasmas are connected. Under these conditions any potential applied to the anode in excess of the arc drop will cause refiring. From (1) and (3) we obtain

$$T = \tau \ln \frac{i_{g0}R_g}{E_g}. \tag{4}$$

From (4) we see that the greater $i_{g0}$ and $R_g$, the longer will recovery be delayed. This accounts for the effect of $R_g$ on $T_R$. Furthermore, we see that if $i_{g0}$ can be decreased, then it becomes possible to get more rapid recovery. Incidentally, if $i_{g0}$ can be decreased sufficiently, then it is possible to increase $R_g$ and still obtain reduced $T_R$. But increased $R_g$ means increased sensitivity to signal current, a desirable result. A means for decreasing $i_{g0}$ is to decrease the effective area of the grid. We shall see later how this can be effected by a separation of the firing and recontrol functions of the grid.

## CONSIDERATIONS REGARDING GRID EMISSION

It has been shown that $T_R$ can be diminished by decreasing the interelectrode spacings. However, when an effort is made to go in this direction, one generally runs into difficulties with increased grid emission due to increased grid temperature, particularly if power levels are not reduced. This grid emission, when flowing through the grid resistor, results in a reduction of the effective grid bias. Since the emission may be nonconstant in time, it can lead to erratic performance, in that the bias may need frequent readjustment in order to prevent a continuous discharge.

Temperature measurements were made on a conventional thyratron triode (see Fig. 4), by means of thermocouples mounted on the grid and anode. Fig. 5 shows the temperature of the electrodes as the tube warmed up and was fired. From the temperature data and from grid-emission measurements, the work function of the grid surface was found to be approximately 1.5 volts. The constant $A$ of Richardson's equation was 5 amperes per square centimeter per degree squared. During normal operation of the tube, the grid reaches a temperature of 600°K. The close spacings and the presence of an oxide cathode with its accompanying barium evaporation onto the grid make it appear that grid emission would not be unexpected. Fig. 6 shows curves of grid emission from the commercial thyratron (of the form shown in Fig. 4) under various operating conditions. In these tests the grid was held at −25 volts with respect to cathode and anode during the reading of the grid current. This value of potential served to saturate the grid emission without introducing undue ionization in the gas. In the cases where plate current was drawn, the grid emission was measured while the plate was momentarily shorted to ground. The emission variations (particularly the decrease in the case where $E_f = 6.3$ v, $I_b = 150$ ma) indicate that besides temperature variation of the grid, the barium distribution over its surface must have been changing with consequent change in effective thermionic constants. It is obvious from these figures that if a large grid resistor were employed the bias required to just hold off firing would change with time, thus resulting in unstable operation.

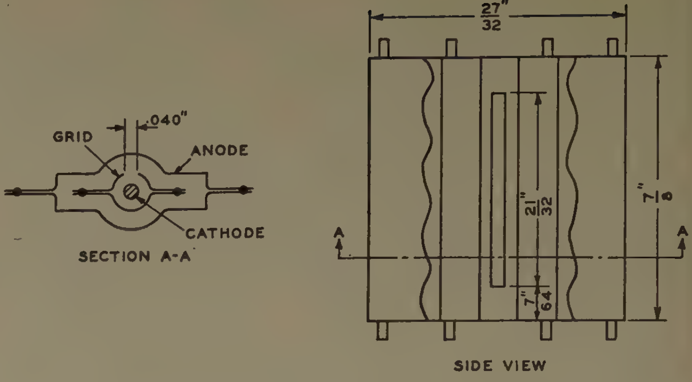

Fig. 4—Sections through conventional thyratron.

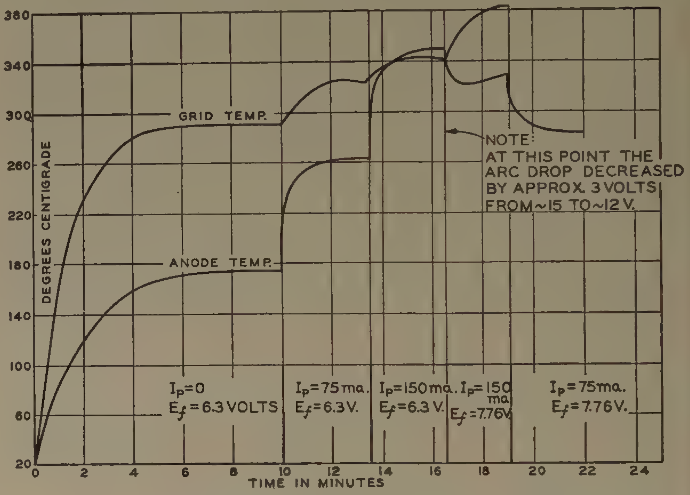

Fig. 5—Thyratron grid and anode temperatures following power turn-on.

Another factor which makes the effects of grid emission serious is that the electric fields may be sufficient to produce field intensified ionization in the gas. Any negative grid bias or positive anode voltage will serve to accelerate emitted electrons and produce ionization which tends to increase the grid current even more. Changes in grid current by factors of over 20 are observed when the grid potential is varied between 0 and −200 volts.

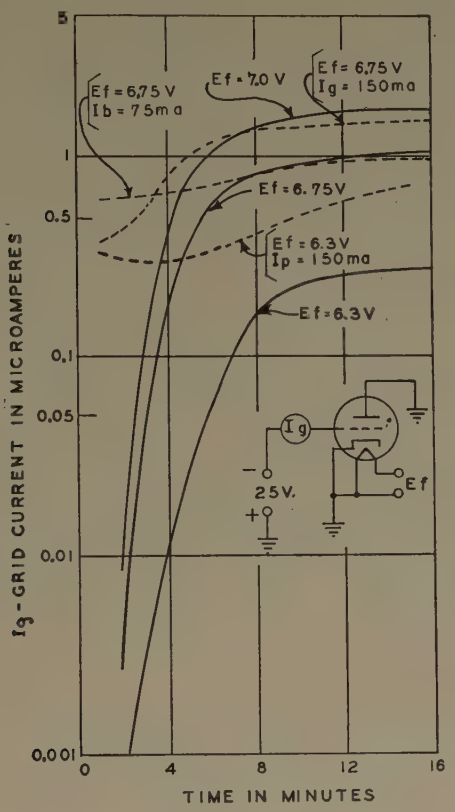

Fig. 6—Thyratron grid current following power turn-on.

## RECAPITULATION AND PROBLEM FORMULATION

It appears that two obvious methods of securing reduced $T_R$ are:

1. Decreasing electrode spacings
2. Reducing positive-ion current to control grid during the afterglow.

If the first of these methods is adopted, grid emission increases with resultant reduction in stability of the control characteristics. This can be overcome by reduction of effective grid area or of grid resistance. Decreasing the size of the grid resistance is undesirable in that it results in a reduced sensitivity to signal current. But the reduction of effective grid area is just what method 2 calls for in order to secure reduced $T_R$. Thus the indications are that the direction of attack should be to reduce spacings and to decrease substantially the effective area of the control grid.

The grid in the conventional three-electrode thyratron has a double function: (1) it serves to initiate the discharge, and (2) it serves to produce a sheath of sufficient thickness during the afterglow, to isolate the plasmas on the plate and cathode sides, thus making recovery of grid control possible. A separation of these functions between two electrodes can be of advantage in that the firing electrode, which is the one whose electron emission and whose positive-ion collection during the afterglow are deleterious, can be made small in order to minimize these effects.

## THE TRIGGER-GRID THYRATRON—DESIGN AND PRINCIPLES

On the basis of the preceding observations and considerations an experimental thyratron of the form shown in Fig. 7 was developed. A typical operating circuit is illustrated in the same figure. It is seen that this modified thyratron differs from that of the standard commercial form shown in Fig. 4, in that an additional electrode in the form of a fine wire has been added and placed within the widened grid opening of the former structure. The fine wire is referred to as a trigger grid because of the fact that its major function is to trigger the discharge. The tube is designated as a "trigger-grid thyratron" (TGT). The recovery function which is normally played by the same grid that initiates the discharge, is now achieved primarily by the blocking grid. The input signal is applied to the trigger grid only. Because of the fineness of the trigger grid, the positive-ion current to it during the afterglow is very small. As a consequence, even though $R_g$ is made large, the trigger grid can return to its bias value quite rapidly, thus giving reduced values of $T_R$. Concomitantly, one achieves an improved signal current sensitivity due to the permissibly larger value of $R_g$. Since the blocking grid resistor $R_L$ need be only large enough to prevent the flow of excessive grid currents during the discharge, the blocking grid returns to its bias potential almost instantaneously following the interruption of the discharge. (It was found that a value of $R_L$ of about 5,000 ohms worked satisfactorily in the experimental tubes.) The consequent very rapid sheath buildup around the blocking grid aperture promotes rapid recovery. It is

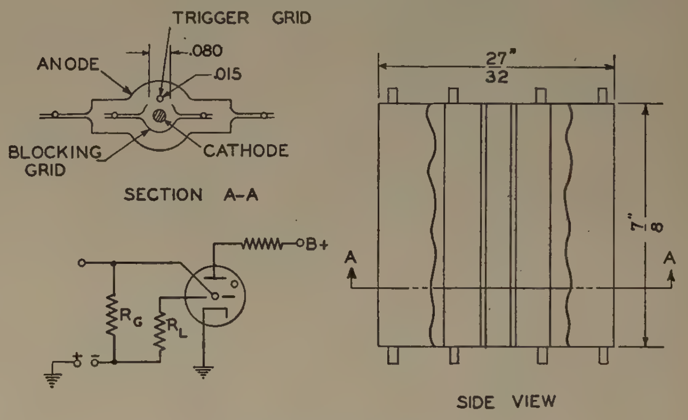

Fig. 7—Sections through trigger-grid thyratron. TGT operating circuit.

seen thus, that by the incorporation of a trigger grid into the structure of a three-electrode thyratron so as to achieve a separation of functions of firing and sheath formation, one profits by the reduced current flow to the trigger grid during the afterglow.

Another advantage of this design is in the reduced grid-emission effects. Since the blocking grid is tied to the bias source through a low-value resistor, the potential drop across this resistor due to grid emission is negligible compared to the bias potential. Thus the unstable characteristics, which can exist when the triggering and blocking functions are combined in one grid, are obviated. This will be brought out in the results presented in the following section.

## THE TRIGGER-GRID THYRATRON—EXPERIMENTAL RESULTS

The advantages of the TGT are brought out in comparative studies. These included measurement of the following:

A. Grid emission
B. Control characteristics
C. Critical grid bias
D. Recovery characteristics
E. Interelectrode capacitance
F. Arc drop.

The results will be discussed seriatim.

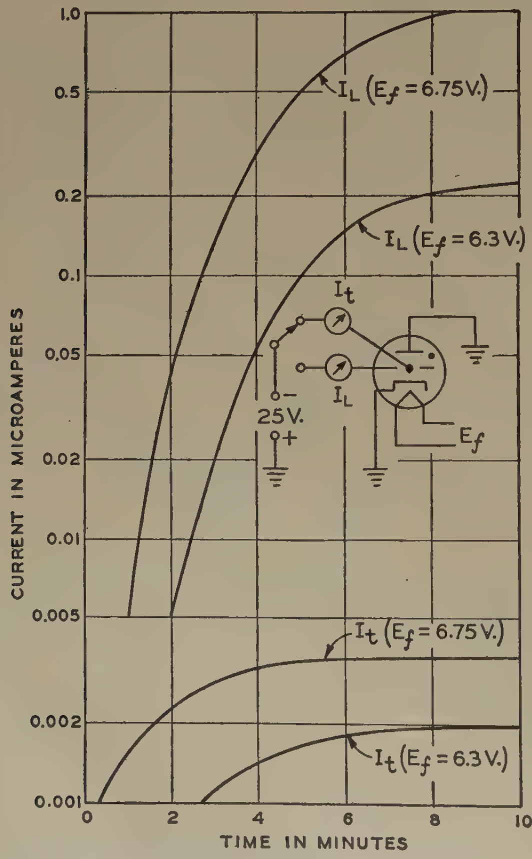

Fig. 8—Trigger-grid and blocking-grid currents following power turn-on.

### A. Grid Emission

Grid emission was measured in the circuit shown in Fig. 8. The emission from trigger and blocking grids as a function of time following turning on the heater is shown in Fig. 8. It is seen that assigning the firing action to a small-area electrode has resulted in marked reduction in grid emission from the electrode of importance—the electrode which is connected to the input signal resistor. A comparison of Fig. 8 with Fig. 6 shows that the emission from the blocking grid of the TGT is about the same as that from the control grid of the conventional triode. Thus the grid emission, in so far as its effect on control characteristics is concerned, is seen to have been reduced by a factor well in excess of 100.

Cases for which plate current was drawn were also studied. It was found, as above, that the trigger-grid emission is of the order of 40 to 100 times less than that from the blocking grid of the same tube or that from the control grid of its conventional prototype. (In making these measurements the grid emission was determined while the anode was momentarily shorted to the cathode.)

### B. Control Characteristics

Since the TGT contains two grids, a complete presentation of its control characteristics (CC) requires a family of curves. These are presented in Fig. 9. This figure also includes the control characteristic of the conventional thyratron of Fig. 4. A comparison indicates

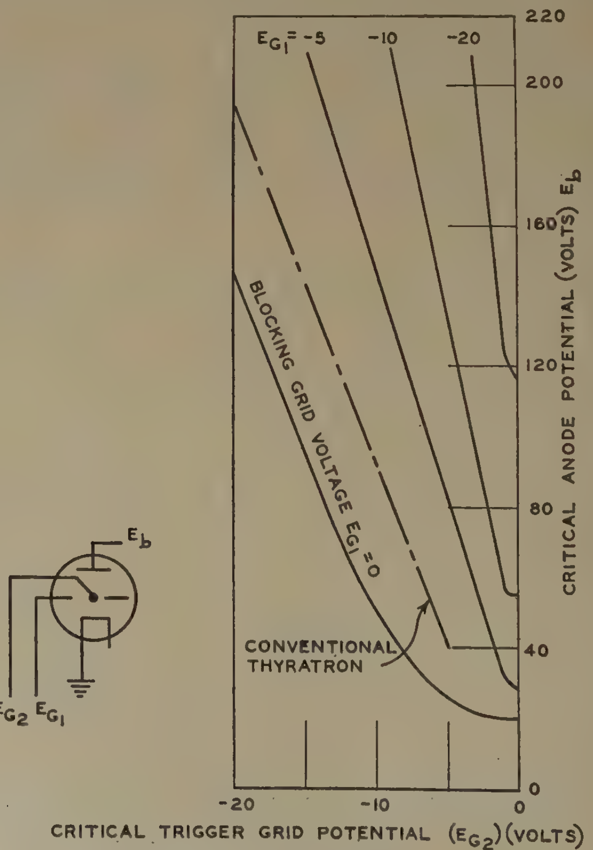

Fig. 9—TGT control characteristics.

that the experimental TGT has a somewhat higher control ratio than the conventional thyratron. Actually, by changes in the size of the blocking-grid aperture and of the position of the trigger grid, the control ratio of the TGT can be varied over a wide range around that of the conventional thyratron of Fig. 4.

### C. Critical Grid Bias

The critical grid voltage (CGV) is the grid voltage at which the tube just fires for a particular value of anode voltage. Values of this quantity are given in Fig. 9. If there is no grid emission, then the bias source potential corresponding to that state at which the tube just fires is equal to the CGV. When grid emission is present, then, due to the drop in the grid resistor, the bias potential must be increased so as to prevent firing. We shall call this value of bias potential, the "critical grid bias" (CGB). For dc potentials on the anode, the CGB can be computed from the values of CGV, grid emission, and the magnitude of the grid resistor. If the anode signal is ac in character, then due to capacitance between anode and grid, a portion of the anode signal will appear at the grid (when a grid resistor is present). A still greater bias will then be required to prevent firing.

A set of data to show the relative merits of the various tube types for the case in which the anode voltage is ac in nature is shown in Fig. 10. In this case, instead of operating at fixed values of anode voltage and grid resistance, critical bias values over a range of anode voltage were determined for a number of values of grid resistance. The resultant curves bear a resemblance to the control characteristics and, in fact, coincide with them when $R_g=0$. In taking these data, the tube was permitted to "settle down" to a final steady value of CGB before recording the data. The curves include within themselves not only the effect of primary grid emission, but of gas amplification and of the voltage induced on the grid as a consequence of the interelectrode capacitance. The curves need no comment other than that they, too, clearly show the decidedly greater stability of the TGT, in this case with ac on the plate.

### D. Recovery Characteristics

The recovery characteristics of a TGT and of the conventional thyratron of Fig. 4 are shown in Fig. 11.

In these measurements, the tube was fired for 6 microseconds at an anode current of 250 ma. A later 200-volt test pulse was applied to the anode to determine whether the grid had recovered control. The $T_R$ characteristic for the conventional thyratron of Fig. 4 is not very different from the curve obtained by joining the points of the TGT data for which $V_{G1}=V_{G2}$. This is to be expected since, from the point of view of the

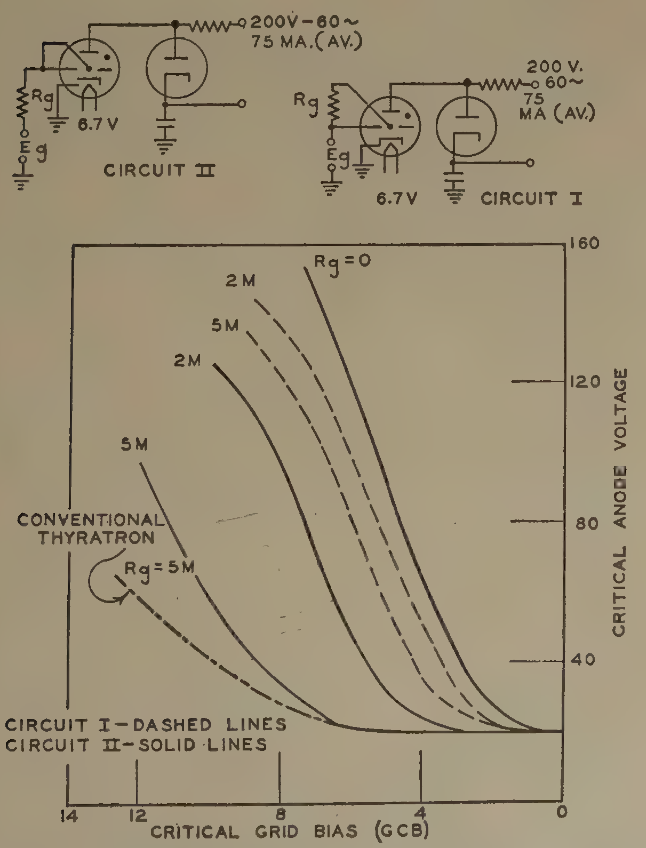

Fig. 10—Control characteristics of TGT and conventional thyratrons for various values of grid resistor.

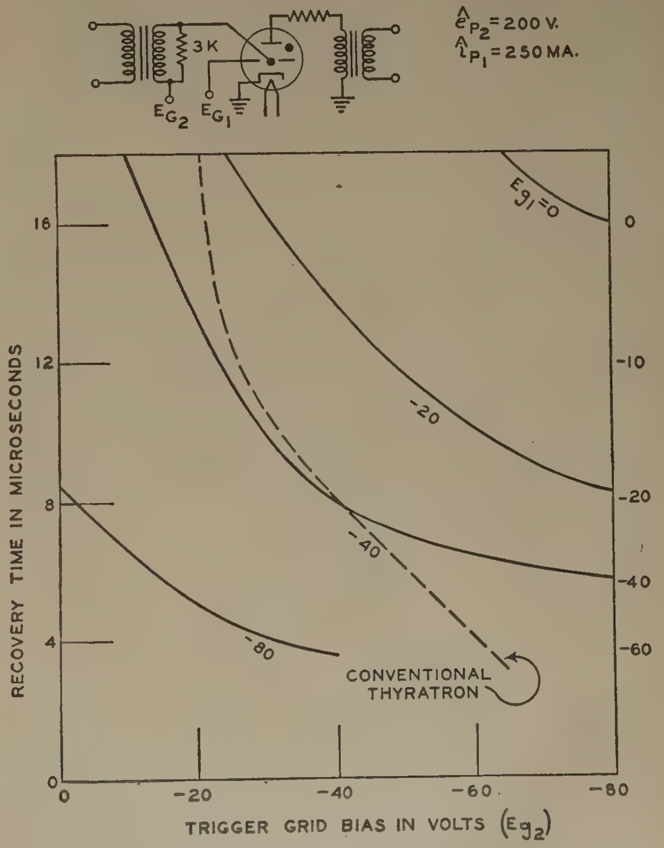

Fig. 11—Recovery times of TGT and conventional thyratron as function of grid voltage.

recovery mechanism, the TGT tube with joined trigger and blocking grids is similar to the prototype thyraton.

As was explained above, the effect of grid resistance is to prevent the applied dc bias from appearing at the grid immediately after the interruption of the discharge because of the voltage drop due to positive-ion current flowing to the grid and thus through the grid resistance. This decreased effective bias results in a longer $T_R$. The magnitude of the positive-ion current from the decaying plasma to any electrode is almost independent of its potential, but is very much dependent on its area. Thus, if the positive-ion collecting area is made small, the detrimental effect of grid resistance on $T_R$ is reduced. The trigger electrode, designed with a small area to minimize emission, therefore, has the property of collecting less ion current during the decay period. Consequently, it returns to the bias voltage value more rapidly. Fig. 12 brings this out and shows the superiority of the TGT with respect to $T_R$ when series grid resistance is present. The observed behavior would be advantageous in many applications where low $T_R$ is desired.

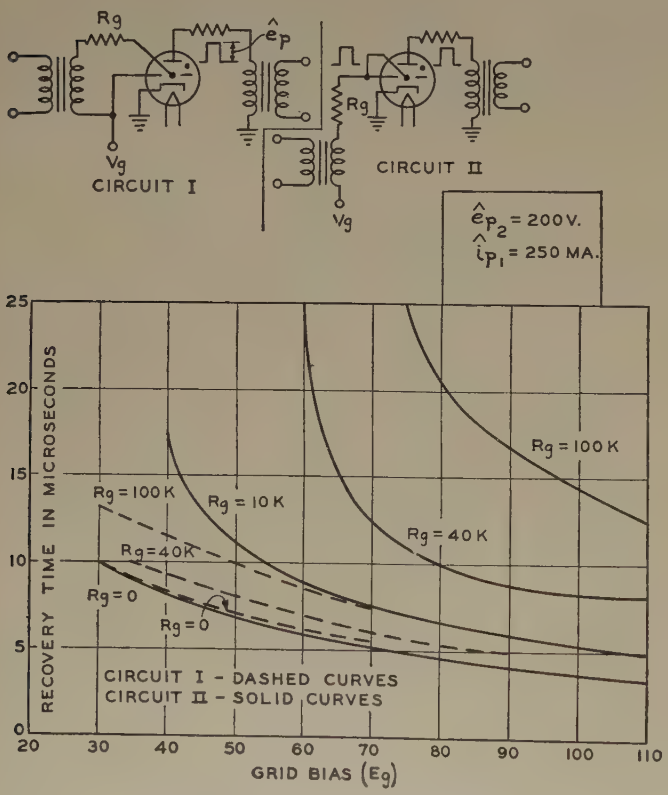

Fig. 12—Recovery times of TGT for various values of grid resistor.

### E. Interelectrode Capacitance

In Fig. 12, part of the observed superiority of the TG tube is due to the much lower capacitance between the TG and the anode than between the control grid and anode of the conventional thyratron. The reduced capacitance results in a reduction of the signal induced on the control electrode by varying anode voltage. Thus in the case of the curves of Fig. 12, the instantaneous grid voltage values may be appreciably less than the bias values.

The interelectrode capacitances were measured on a bridge with the results shown in Table I.

TABLE I

| Electrodes | Capacitance |
|---|---|
| Trigger grid to anode | 0.83 $\mu\mu$f |
| Trigger grid to cathode | 1.11 $\mu\mu$f |
| Trigger grid to blocking grid | 0.77 $\mu\mu$f |
| Blocking grid to anode | 5.06 $\mu\mu$f |
| Blocking grid to cathode | 1.61 $\mu\mu$f |

These results should be compared with those for the conventional thyratron of Fig. 4. The rated values for this tube are: grid to anode capacitance $=6$ $\mu\mu f$; grid to cathode capacitance $=2$ $\mu\mu f$.

Thus, there is essentially no difference in the input capacitance, but there is a decrease in the feedback capacitance by a factor of about seven. This is very helpful at high-frequency operation, particularly with high grid resistance, since it means that a much smaller disturbing signal appears on the control electrode due to the presence of a varying voltage on the anode. As was seen above, this makes itself felt in smaller effective $T_R$.

### F. Arc Drop

Arc drop readings were taken on the TGT on the thyratron of Fig. 4 and found to be substantially identical.

## SUMMARY COMPARISON OF THE TRIGGER-GRID AND CONVENTIONAL THYRATRONS

From the experiments described above it is apparent that the TGT has a number of advantages over its conventional prototype particularly in the presence of a large grid resistance.

A summary comparison between permissible operating values of the experimental TGT and its conventional prototype is given in Table II.

TABLE II

| Characteristic | Conventional Thyratron | TG Tube |
|---|---|---|
| Grid emission<br>$R_g$ max<br>$I_p$ max<br>$E_f$ max | 1–10 $\mu$amps<br>0.5 mc*<br>75 ma*<br>6.7 v | ~0.05 $\mu$amps<br>> 10 mc<br>>200 ma<br>>7.7 v |
| Recovery time | | Lower than conventional thyratron when $R_g \neq 0$ |
| $C_{g-p}$ | 6 $\mu\mu$f\* | 0.8 $\mu\mu$f |
| Control characteristics | | Similar to conventional thyratron when $R_g \approx 0$<br>Improved when $R_g \neq 0$ |

* Rated values.

## SUMMARY AND CONCLUSIONS

It has been shown that the limitations on a thyratron are set in part by grid emission, and in part by positive-ion-grid current during the afterglow. These factors set an upper limit to the permissible value of grid resistor and plate current. Exceeding these limits results in unstable and time-varying control characteristics and in delayed recovery of grid control.

Efforts to reduce recovery time by reducing electrode spacing lead to increased grid emission with consequent unstable performance.

A promising means for minimizing the limitations involves adding to modified conventional structures an additional small area electrode which is used as the control or firing electrode. Tubes incorporating this extra electrode are known as trigger-grid thyratrons. The grid of a conventional thyratron normally serves a double function, i.e., of firing the tube and of promoting recovery of grid control following the interruption of a discharge. In the TGT these functions are largely separated. The firing function is assigned to the trigger grid only, while the recovery is accomplished mainly by the action of the so-called blocking grid. The benefits of this design are due in part to the reduced emission from the small area of the trigger grid and in part to its smaller collecting area for positive ions during the afterglow. On test it was found that the current rating of an experimental TGT tube was superior to its conventional prototype as regards permissible plate current, grid resistance, and heater voltage. The TGT is also superior as regards recovery characteristics and the effects of grid-plate capacitance.

While the experimental studies and development centered around a low-power thyratron, the principles involved and the methods developed appear applicable to thyratrons of any power rating.

### ACKNOWLEDGMENT

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# A Study of Tropospheric Scattering of Radio Waves*

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***Summary*—The theory of tropospheric scattering of radio waves recently advanced by Booker and Gordon[1] makes the prediction that radio-frequency energy may arrive from a ground-based very-high-frequency transmitter at angles well above the horizon, and also that signals received as a result of the scattering process will attenuate less rapidly with distance if the polarization of the source is horizontal than if it is vertical. Experimental observations made at wavelengths of 3 meters and 3 centimeters have verified these deductions.**

## I. INTRODUCTION

RADIO FIELD INTENSITIES observed at distances considerably beyond the line of sight have regularly been found to be much higher than would be expected on the basis of the conventional refraction theory of tropospheric waves. This has been noted for both overwater and overland propagation,[2–4] but until this year no satisfactory theoretical treatment has been brought forth.

Booker and Gordon[1] have recently applied the theory of scattering to radio waves in the troposphere. From their results, they conclude that a radio wave beyond the horizon of the transmitting source is made up of a component due to refraction and also a component due to scattering, the latter often being as strong as, or stronger than, the refraction component. At distances not far beyond the horizon, the refracted wave is predominant and the scattered wave is then a cause of fading.

At the greater distances where the result of scattering predominates, the scattering theory predicts that a considerable portion of the energy will arrive from angles well above the horizon. One result of this is an apparent broadening of the acceptance pattern of a high-gain antenna system. The theory further predicts that at great ranges the attenuation of signal with distance is greater for vertical than for horizontal polarization.

A number of experiments performed during the past year at the Electrical Engineering Research Laboratory of The University of Texas have largely verified these predictions. This paper will describe these experiments and present some of the findings.

## II. DISTRIBUTION OF SCATTERING FROM A TURBULENT REGION

Booker and Gordon give the scattered power $\sigma(\theta, x)$ per unit solid angle, per unit incident power density, and per unit macroscopic element of volume by

$$\sigma(\theta, x) = \frac{\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2} \left(\frac{2\pi l}{\lambda}\right)^3 \sin^2 x}{\lambda\left[1 + \left(\frac{4\pi l}{\lambda}\sin\frac{\theta}{2}\right)^2\right]^2} \tag{1}$$

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[2] J. P. Day and L. G. Trolese, "Propagation of short radio waves over desert terrain," PROC. I.R.E., vol. 38, pp. 165–175; February, 1950.

[3] M. Katzin, R. W. Bauchman, and W. Binnian, "3- and 9-centimeter propagation in low ocean ducts," PROC. I.R.E., vol. 35, pp. 891–905; September, 1947.

[4] K. A. Norton, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. 1, pp. 381–423; 1948.

where

$\epsilon$ = the average permittivity
$\Delta\epsilon$ = the departure of the permittivity from its average value
$l$ = the scale of turbulence
$\lambda$ = the wavelength
$x$ = the angle between the direction of the electric field vector and the direction to the receiver
$\theta$ = the angle between a line drawn from the transmitter through the scattering region and a line drawn from the scattering region to the receiver. ($\theta$ is small when the scattering region is near the optical horizon).

Referring to Fig. 1, for horizontal polarization the value of angle $x$ will be 90° when the plane formed by the lines from the scattering point to the transmitter and to the receiver is normal to the earth. The sine of $x$ will therefore be approximately unity with the scattering region located anywhere between overhead and the horizon.

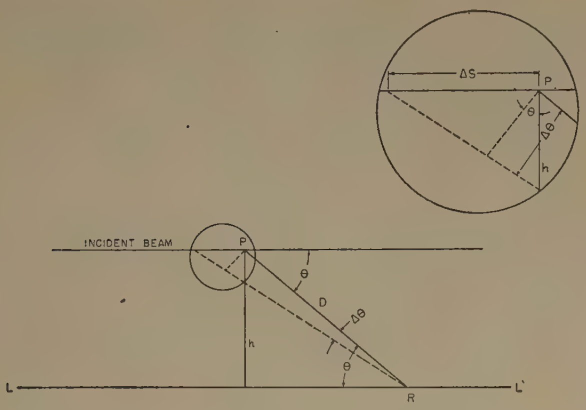

Fig. 1—Geometric quantities in scattering formula.

For vertical polarization, however, the angle $x$ will be zero when the scattering region is overhead and will approach 90° when the scattering region nears the horizon. Thus, scattered signals for vertical polarization will be reduced except when arriving horizontally.

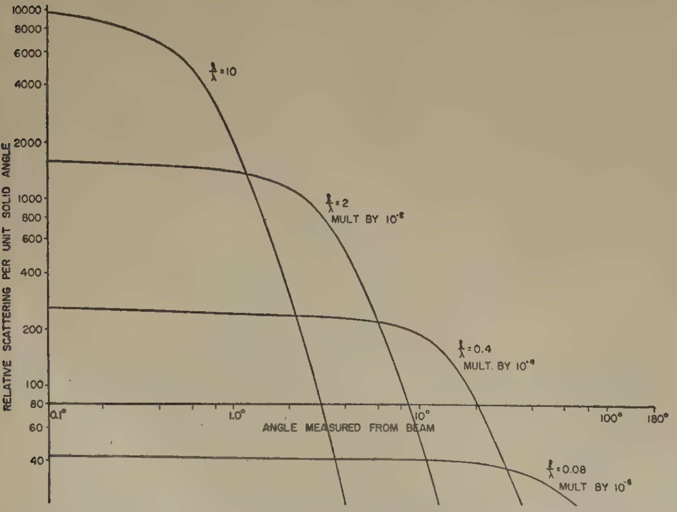

Fig. 2—Relative intensity of scattered signal per unit solid angle, horizontal polarization.

The relative distribution with the angle $\theta$ is shown in Fig. 2 for four values of $l/\lambda$ and for horizontal polarization. If the scale of turbulence is taken as 30 cm, $l/\lambda = 0.08$ corresponds to a frequency of 80 mc, and $l/\lambda = 10$ corresponds to a frequency of 10,000 mc.

## III. DISTRIBUTION OF SIGNAL SCATTERED FROM AN ISOLATED REGION ON A PLANE PARALLEL TO THE INCIDENT BEAM

If the radio scattering occurs from an isolated region $P$ in Fig. 1, the power density received along a line parallel to the incident beam will be given by

$$w(\theta, x) = \sigma(\theta, x) \Big/ \frac{4D^2}{\pi}. \qquad (2)$$

Replacing $D$ by $h/\sin\theta$ and using the value of $\sigma(\theta, x)$ from (1), we have

$$w(\theta, x) = \frac{\pi \overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2} \left(\frac{2\pi l}{\lambda}\right)^3 \sin^2 x \sin^2 \theta}{4\lambda h^2 \left[1 + \left(\frac{4\pi l}{\lambda} \sin \frac{\theta}{2}\right)^2\right]^2}. \qquad (3)$$

It is of interest to note that this power density along the line $LL'$ will be maximum at an angle $\theta_m$ given by

$$\sin^2 \frac{\theta_m}{2} = \frac{1}{2 + \left(\frac{4\pi l}{\lambda}\right)^2}. \qquad (4)$$

For very large values of $l/\lambda$, the power density will be maximum a long distance from the scattering region (i.e., for $\theta \rightarrow 0$), while for very small values of $l/\lambda$, the maximum power density will be directly below the scattering region.

## IV. POWER DENSITY FROM BEAM OF UNIFORM ILLUMINATION

Let us consider the power density at a receiving point $R$ (Fig. 1) due to scattered radiation originating in an equally illuminated beam. The length of the beam $\Delta s$, which will be included in an incremental angle $\Delta\theta$, will be given by $D\Delta\theta/\sin\theta$ or $h\Delta\theta/\sin^2\theta$. The power intensity, $\rho(\theta, x)$ per unit solid angle at $R$ will be given

$$\rho(\theta, x) = w(\theta, x) \frac{\Delta s}{\Delta\theta} = \frac{\pi}{4\lambda h} \frac{\overline{\left(\frac{\Delta\epsilon}{\epsilon}\right)^2} \left(\frac{2\pi l}{\lambda}\right)^3 \sin^2 x}{\left[1 + \left(\frac{4\pi l}{\lambda} \sin \frac{\theta}{2}\right)^2\right]^2}.$$

This distribution of power intensity as a function of the angle $\theta$ has the same form except for a constant factor as (1). The graph of Fig. 2 may therefore be thought of as the distribution of the signal received from scattering along a uniformly illuminated beam as a function of angle of elevation.

For large values of $l/\lambda$, the scattering would be confined to a small angle near the horizon while for smaller values of $l/\lambda$, the scattered radiation will also be observed at higher elevation angles. Evidence of the arrival of radiation at large elevation angles is presented in the following section.

## V. The Effect of Antenna Directivity on the Reception of Distant FM Stations

As a test of the arrival of radio energy at large elevation angles, a comparison was made of the signal received from the same commercial FM station on two antennas. One of these antennas was a horizontal dipole and the other consisted of a pair of horizontal dipoles separated horizontally one-half wavelength and fed out of phase. The double-dipole antenna system had a gain of 6.5 db over the simple dipole. This includes transmission-line losses in both systems, which were different. Radiation at right angles to the main lobe of the double-dipole antenna and in a plane normal to the dipole elements was down 15 db from the radiation level in the direction of the main lobe. Two types of antenna comparisons were made as follows:

(1) Directive antenna pointed toward the FM transmitter.

(2) Directive antenna pointed vertically.

### *A. Directive Antenna Pointed Toward the FM Transmitter.*

Simultaneous recording on the two antennas were made from FM stations in Dallas, Houston, and San Antonio, Texas. These stations have frequencies of 97.9, 102.9, and 101.5 mc, respectively, and are at distances of 182, 146, and 73 miles from the receiving site in Austin. The signal levels which were exceeded for various percentages of time were determined from computers or graphs for each hour recorded.

In comparing the signals from Dallas and Houston delivered by the two antennas to the receiving systems, it was noted that the ratio of the signals received from the two antennas was a function of signal level. This trend is illustrated in Fig. 3 for the 50-per cent signal from Dallas for the period December 13–16, 1949. This was a period following the entrance of a cold air mass into the state during which there was a continuous advection of cold air and synoptic conditions were fairly constant.

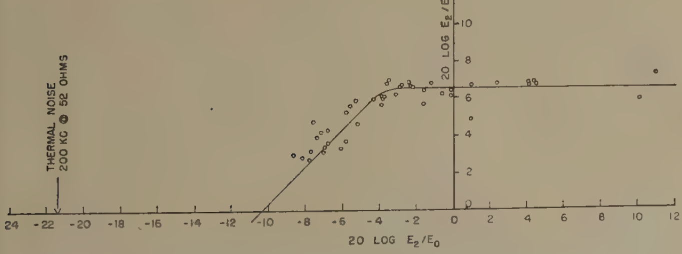

Fig. 3—Comparative signals received with directive and nondirective antennas over 182-mile path, 97.9 mc.
$E_1$=open-circuit voltage, dipole antenna
$E_2$=open-circuit voltage, directive antenna
$E_0$=4.8$(10^{-6})$ volts.

This same trend was seen in all of the data for the 146- and 182-mile paths although the deviation of the measured points from a smooth curve was greater in some cases. This deviation was attributed to changes in meteorological conditions which occurred during the intervals of measurement.

It is noted in Fig. 3 that, for strong signals, the measured ratio was the nominal gain of the directive antenna over the dipole antenna, indicating that the signals were arriving horizontally. For weaker signals, however, the signal ratio in decibels decreased linearly with the strength of the signal received by the directive antenna. This suggests that the weaker signals came from higher elevation angles. The curve shown was drawn through the points in Fig. 3 by eye. The low-signal asymptote of this curve was a straight line with a slope of 45° which was the condition which would exist if the signal from overhead was statistically independent of time as the signal arriving horizontally became very small. The high-level asymptote was the ratio of response which would exist for a signal arriving horizontally. The values of this scattered signal for a number of four-day periods are shown in Table I. Because of the spread of the measured points, these values are subject to a possible error of ±0.5 db.

TABLE I

Scattered Signal Strength and Strength of Refracted Wave as Calculated for a Standard Atmosphere

| Date | Decibel Relative to 1 Microvolt/Meter/Kilowatt | | | Microvolt/Meter/Kilowatt | | |
|---|---|---|---|---|---|---|
| KPRC-FM, Houston—146 Miles | | | | | | |
| Refracted Wave, Calculated for Standard Atmosphere | | −45 | | | 0.0058 | |
| Scattered Signal | per cent 20 | per cent 50 | per cent 80 | per cent 20 | per cent 50 | per cent 80 |
| January 7–10, 1950 | −5.7 | −8.8 | −10.7 | 0.52 | 0.36 | 0.29 |
| January 3–6, 1950 | −5.2 | −7.3 | −9.7 | 0.55 | 0.43 | 0.33 |
| December 30, 1949–January 2, 1950 | −4.8 | −6.0 | −7.0 | 0.58 | 0.50 | 0.45 |
| WFAA-FM, Dallas—182 Miles | | | | | | |
| Refracted Wave, Calculated for Standard Atmosphere | | −61 | | | 0.00084 | |
| Scattered Signal | per cent 20 | per cent 50 | per cent 80 | per cent 20 | per cent 50 | per cent 80 |
| December 17–21, 1949 | −8.0 | −9.3 | −12.0 | 0.40 | 0.34 | 0.25 |
| December 13–16, 1949 | −8.0 | −10.0 | −13.6 | 0.40 | 0.32 | 0.21 |

Table I indicates that the scattered signal was 35 to 50 db stronger than the signal which would have been received by refraction through a standard atmosphere.

The conclusion from these data was that the weaker signals from the distance of 146 and 182 miles came primarily from higher elevation angles and may be attributed to scattering, while the stronger signals were an example of the well-known phenomenon of super-refraction.

For the San Antonio station at 73 miles, the relative signal received on the two antennas indicated that the signal was arriving horizontally. The observed signal was of the same order of magnitude as the refracted signal calculated for a standard atmosphere. The signal from scattering then appeared to be superimposed on the normal refracted signal, but did not change its median value appreciably. For periods of slight fading, the statistical distribution of the signal was Gaussian. However, for periods of deep and rapid fading, the Rayleigh distribution was approximated.

### B. *Directive Antenna Pointed Vertically*

In order to further check the arrival of the FM signals from overhead, the directive antenna was pointed with the minimum in its antenna pattern toward the transmitter and with its main lobe vertical. Simultaneous measurements were made of the signal received from Dallas by the directive antenna in this position and by the dipole antenna. The signal ratio of the antennas was again observed to be a function of the signal strength as shown in Fig. 4. These data were for March 15 and 16, 1950, when the synoptic conditions over Texas were fairly uniform and constant. The ratio of the signal received by the dipole antenna to that received by the directive antenna pointed vertically was plotted as the ordinate while the signal received by the dipole was plotted as the abscissa. The dotted extension of the curve intersects the ordinate axis at 8.5 db, which was the measured ratio of the response of the two antennas to a nearby source from which waves arrived horizontally. For weaker signals, the antenna pointed vertically shows an improvement in signal relative to the dipole. This again indicated that stronger signals which come from the horizon were due to superrefraction while the weaker signals which came in part from overhead were due to scattering.

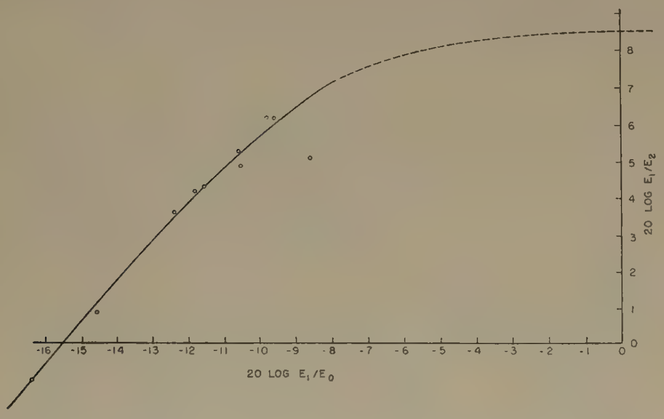

Fig. 4—Comparison of dipole antenna with directive antenna pointed vertically. Station WFAA—FM. Distance, 182 miles. March 15–16, 1950.
$E_1$=open-circuit voltage, dipole antenna
$E_2$=open-circuit voltage, directive antenna
$E_0=4.8(10^{-6})$ volts.

## VI. Effect of Polarization on Reception of FM Signals

The power density at a given point is seen from (3) to be proportional to the square of the sine of the angle between the electric field vector and the direction from the scattering region to the receiver. This factor will cause the scattered signal arriving from overhead to be less for vertical polarization than for horizontal polarization. In addition, a vertically oriented antenna will have a reception pattern which will discriminate against signals arriving from overhead. These factors indicate that the scattered signal for vertical polarization should drop off more rapidly with distance than that for a horizontally polarized antenna when $l/\lambda$ is small.

Two tests of the effect of polarization were made for a frequency of 100 mc. These tests are not conclusive but they do show an indication that the vertically polarized signals are attenuated faster at distances beyond the optical horizon than are the horizontally polarized signals.

### A. *Distance Run on Incidental Vertical Polarization*

The signal from station KMHB in Belton, Texas, was observed at a range of distances up to 90 miles by both horizontal and vertical dipole antennas. The antennas were mounted above a small trailer pulled by a truck in which were located the two receivers. Sufficient vertical polarization accrued from the tower to make possible measurement of vertically polarized signals. The KMHB antenna is 300 feet above immediately surrounding ground, but because of its geographical location, the equivalent height over the road followed was very much less.

In order to be able to reduce the results to a comparable radiated power basis, measurements were started within the line of sight of the transmitter. For

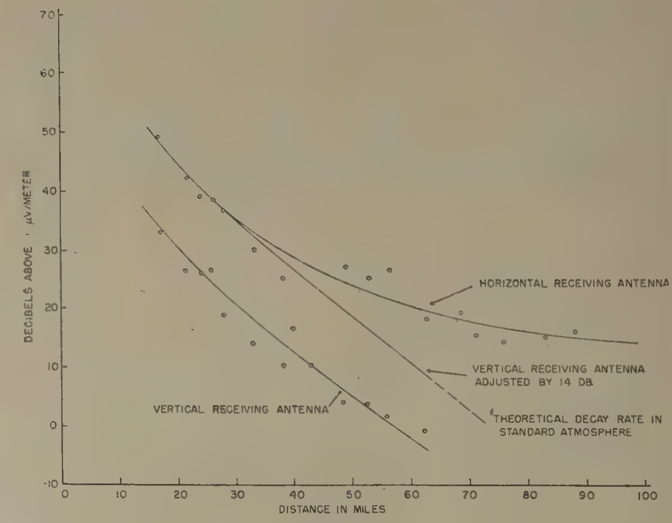

Fig. 5—Comparison of signal attenuation at 97.1 mc for horizontal and vertical polarization. Transmitting source: KMHB, Belton, Texas. Antenna height, 300 feet. Effective radiated power, 12 kw. Receiving antenna height, 10 feet.

distances less than 30 miles, the average signal received by the horizontal antenna was consistently 14 db stronger than the signal received by the vertical antenna. The signal strengths as measured are shown in Fig. 5. In addition, the signal curve for vertical polarization has been redrawn with an increase of 14 db to superimpose it on the curve for horizontal polarization at distances less than 30 miles. The vertical polarization curve was terminated at the distance at which the difference in the signal levels of the two systems approached the known discrimination level of 24 db. From this curve, it is seen that the signal of the vertical antenna decreases at the same rate as the standard refracted wave curve while the signal received by the horizontal antenna decreases more slowly than the theoretical curve. Cross feed between the systems would tend to obscure this difference rather than accentuate it.

### *B. Distance Run on Antenna Polarized at 45°*

As a further test of the polarization, an antenna was polarized at 45° and 100-mc power was radiated from it. The distance run was then repeated. Because of the reduced antenna gain and since the antenna in this case was only 60 feet above surrounding ground, it was not possible to carry this test beyond 30 miles. It was noted that for distances less than 23 miles, a single curve represented both sets of data, while for greater distances, the horizontally polarized component of the signal was somewhat greater than the vertically polarized component.

## VII. FM SIGNALS AS A FUNCTION OF ANGLE OF ELEVATION

In order to further test the variation of scattered signal with elevation angle, a special antenna was constructed which could be tilted through 180°. This antenna was a V-type reflector with planes 20 feet square separated at an angle of 45°. The measured antenna pattern for 100 mc is shown by curve (2) in Fig. 6.

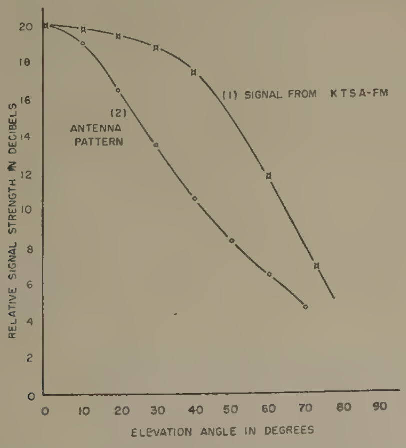

Fig. 6—FM signal versus elevation angle. Distance, 73 miles. Frequency, 101.5 mc. Polarization, horizontal.

Using this antenna at various tilt angles, field-strength recordings were made on the signal received from radio station KTSA-FM in San Antonio, Texas, at a distance of 73 miles. The frequency of this station is 101.5 mc. A 3-minute median level was determined at each tilt angle and the results are plotted as curve (1) in Fig. 6. The effect of diurnal changes in field strength taking place during the experiment was eliminated by means of an auxiliary monitoring receiver using a simple dipole antenna situated near the larger receiving antenna.

It is seen that when scanning the sky for the 100-mc signal, a much flatter curve was obtained than when scanning a nearby point source. This indicates that the radiation is arriving over a range of elevation angles considerably above the horizon.

## VIII. MICROWAVE SIGNALS AS A FUNCTION OF ANGLE OF ELEVATION

It is indicated in Fig. 2 that at wavelengths of a few centimeters and for a scale of turbulence of the order of a meter the energy scattered by the atmosphere will be confined within a few degrees of the horizontal direction. A field test at a wavelength of 3.2 cm was made in the vicinity of Austin. The 20-foot parabola used by the Bell Telephone Laboratories for their original 3-cm angle-of-arrival studies[5] was used for scanning. The center of this parabola was approximately 15 feet above ground, and the transmitting antenna 37 miles away was 12 feet above ground. A pulse signal was used with a peak power of 19 kw. Polarization was horizontal.

Records of the signal received were made in 0.1° steps, and the statistical time distribution of the signal

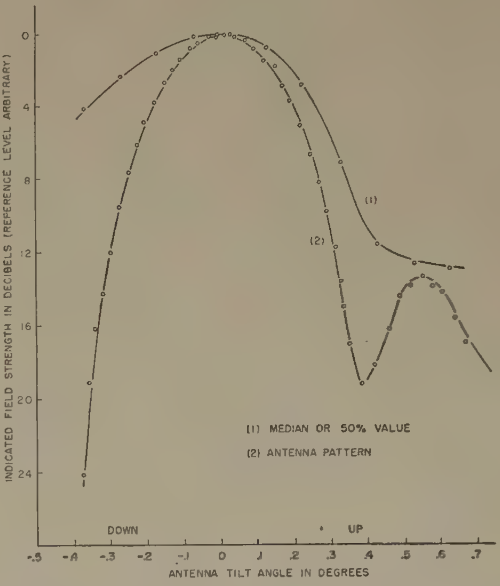

Fig. 7—Microwave signal versus antenna tilt angle. Transmitting antenna: height above ground, 12 feet; aperture, 24λ high×16λ wide. Receiving antenna: height above ground, 15 feet; aperture, 190λ high×5λ wide. Wavelength, 3.2 cm. Polarization, horizontal. Path length, 37 miles.

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[5] W. M. Sharpless, "Measurement of the angle of arrival of microwaves," PROC. I.R.E., vol. 34, pp. 837–845; November, 1946.

determined at each angle. From these distributions, the median value was established and this is plotted on Fig. 7 as a function of the angle of tilt. The antenna pattern is also shown in Fig. 7 for comparison. As in the case of the 100-mc signal, the response distribution is seen to be considerably broader than the antenna pattern.

### IX. Conclusions

A. Comparison of the reception at large distances with a dipole and a more directional antenna at a frequency of 100 mc indicated that:

1. During periods of stronger signals, the apparent source was near the horizon.
2. During periods of weaker signals the apparent source was more diffuse, and considerable radiation was received at higher elevation angles.

B. Distance runs on a 100-mc source indicated an attenuation with distance as given by the refracted wave theory for vertical polarization, but a smaller attenuation for horizontal polarization.

C. When a 100-mc source was scanned in a vertical plane, a distinct broadening of the antenna response was noted. When a 3.2-cm source was scanned, a similar effect was noted which, however, was confined to a few degrees above the horizon.

D. The collection of results of these experiments on tropospheric scattering appears to substantiate the theory of scattering discussed earlier in this paper.

# The Component Theory of Calculating Radio Spectra with Special Reference to Frequency Modulation*

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*Summary*—**A new theory for the resolution of a modulated signal into its radio-frequency spectrum is presented. It is shown that this theory is mathematically equivalent to previous methods of analysis. The component theory introduces physical concepts which give an understanding of the process of formation of sidebands. Examples are worked using an analytical method of analysis. The theory also naturally leads to a graphical method which is applicable without harmonic analysis of the modulation. The concepts apply without modification to any type of modulation although special emphasis is given to frequency modulation.**

## Introduction

A MAJOR difficulty in the analysis of modulation systems has been the transition from the instantaneous frequency concept to the spectrum solution. The "component" method of analysis, described in this paper, supplies a transition and is shown to give results equivalent to those obtained by the usual methods. It is pointed out that the mathematical procedure for obtaining the spectrum components provides a definition of the carrier frequency of an FM wave.

## The Component Method of Analysis

It is of interest to first consider the situation of Fig. 1 which represents a filter with a passband outside the instantaneous frequency range of an FM signal. That energy can be transmitted by the filter is difficult to perceive. Equally difficult to understand is how no energy would be transmitted by a filter which passes part of the instantaneous frequency range of the signal. The trouble lies in considering the usual filter steady-state frequency response as being applicable to this situation. Instead, we must treat the experiment from a pseudo-transient point of view. The varying frequency voltage at any instant will cause a current to flow in the filter network, or it will affect the current which is flowing as a result of the voltage applied during the preceding small time interval. If the voltage at any time has a component in phase with the current existing in the filter, then energy will be added to the filter. If the voltage has a component out of phase, energy will be abstracted from the filter. The net energy transfer from the applied signal to the filter is obtained by averaging over a time equal to one period of this process. The existence of a spectrum component is dependent therefore on the nonvanishing of the average energy transfer. This approach is applicable to any type of modulated signal.

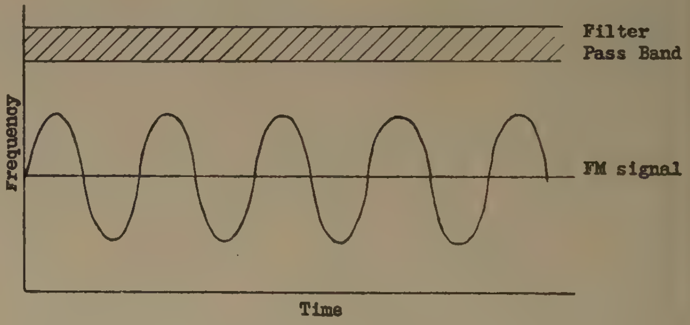

Fig. 1—Example of an FM signal applied to a narrow-band filter.

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To develop these ideas, consider a graphical representation of an FM signal as in Fig. 2. The length of the directed line segment is equal to the signal amplitude and the speed of rotation of the directed line segment about the origin is equal to the instantaneous angular frequency of the signal in radians per second. Such a directed line segment is called a phasor. In general, the phase angle of an FM signal is always *monotonically* increasing, but at certain portions of the modulation cycle the rate of increase is greater than at other portions. If the plane of the figure upon which the phasor is drawn is considered to be rotating in a positive sense (i.e., counterclockwise) at the carrier angular frequency (carrying the observer with it), then the phasor representing the FM signal will be seen to swing alternately clockwise and counterclockwise, where the amplitude of the angular swing may include many rotations. Thus if the carrier signal is considered without angular modulation, its speed of rotation is equal to that of the plane upon which it is drawn, and the phasor remains stationary at point *A*. When angular modulation is applied, the phasor may increase its angular velocity corresponding to the increase in frequency of the signal and will move to *B*, then reverse its motion to *C* as the signal frequency decreases, then repeat the cycle. The phasor amplitude is constant, indicating the absence of amplitude modulation.

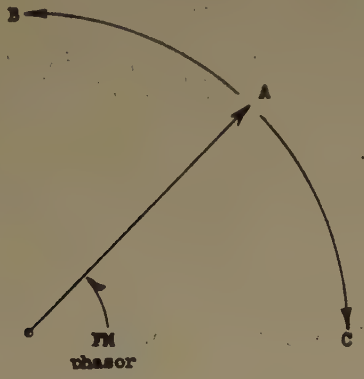


Fig. 2—Graphical representation of an FM signal.

If the speed of rotation of the plane were other than the carrier angular velocity, the signal phasor would have a continuous rotational velocity in one direction and superimposed upon this motion would be the alternating motion due to the frequency or phase modulation. It is seen that this picture presents an approach to the definition of the carrier frequency as that frequency with which it is necessary to consider the plane rotating so that the FM phasor possesses only alternating motion. In other words, the positive and negative phase angle swings shall be equal, corresponding to the mathematical definition of "equal areas."

In the following, there will be no sharp distinction between the abstract picture of the spectrum components of a signal and the experimental procedure which produces these spectrum frequencies, as for example, by means of a narrow bandpass filter. Thus, consider a filter of resonant angular frequency Ω excited by a frequency-modulated signal voltage. Upon the plane of Fig. 3 represent the FM signal and the response voltage at the natural frequency of the filter by phasors. If the plane is rotating at a frequency Ω, then the phasor representing the filter response voltage is stationary, while the FM signal phasor will be in motion as a result of its modulation and since in general its carrier frequency is different from Ω.

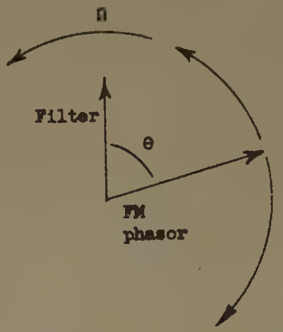


Fig. 3—Graphical representation of signal and filter voltages.

Assume that the component of the FM signal phasor along the direction of the phasor representing the filter response represents the effective voltage to which the filter responds. A current will flow as a result which is proportional to the magnitude of this voltage component, i.e., proportional to the magnitude of the FM signal phasor multiplied by cosine $\theta$, where $\theta$ is the instantaneous angle between the signal phasor and the filter response phasor. It is possible to discuss this assumption in such a way as to make it seem quite plausible, but in the final analysis the validity of such an assumption must be justified by the validity of the results which follow. The results will be seen to be correct in all cases. For values of $\theta$ less than 90 degrees, the component of the signal voltage along the direction of the voltage response phasor corresponds to energy transfer to the filter. The cosine $\theta$ is seen to behave like a power factor. Considering the steady state which will be attained, it appears that the phase of the current in the filter will adjust itself in such a way as to abstract a maximum of energy from the forcing voltage. A useful analogy is to consider how a child's swing adjusts its phase relative to a periodic forcing push so as to reach a maximum height. Hence the phasor representing the filter response may differ from the indicated position in Fig. 3 by some constant angle. In order to locate the exact position of the filter response phasor with respect to the FM signal phasor, the filter voltage can be represented by a pair of orthogonal phasors both rotating at the same frequency Ω, as in Fig. 4. The response to the forcing voltage is obtained by resolving the FM signal phasor upon the two orthogonal filter phasors. The vectorial resultant of the steady-state responses along these two directions gives the resultant response. Thus, the time average of the components along the two orthogonal filter response phasors is calculated by integration over one period of the modulation and the absolute value of the vectorial resultant is the filter response. This procedure will be illustrated below by a series of examples.

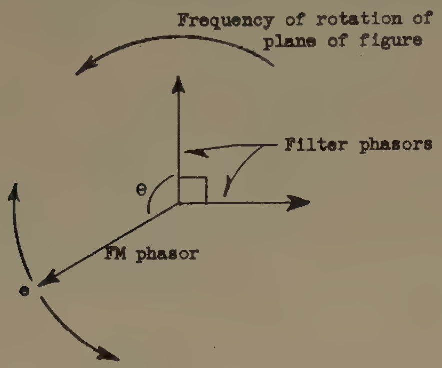


Fig. 4—General representation of signal and filter voltages.

It seems clear at this point that any instantaneous measurement of the filter response (and here a number of pertinent questions as to the meaning of instantaneous measurement are neglected) would indicate a current existing in any filter of any resonant frequency whatsoever. From the view presented here, the forcing voltage component is at one instant in phase with the current, and at a later time opposed in phase to the current. This corresponds to energy alternately being put into and then removed from the filter. In any actual spectrum analysis measurements, the time average of the response of a filter is measured. It must therefore be shown that the time averages of the signal-to-filter energy transfers, computed as explained above, will vanish when the filter does not pass any frequencies corresponding to spectrum lines, and will have the correct nonzero magnitude when the filter passband contains a spectrum component. Several simple examples will be demonstrated, and the general FM spectrum derived.

*Example 1:* Consider an FM signal with simple sinusoidal modulation, and let this signal be applied to a filter containing the carrier frequency $\omega_c$ in its passband.

The input voltage is given by $e = E\cos(\omega_c t + \phi + \beta\sin\omega_m t)$ where

$\omega_c$ = the carrier angular frequency
$\omega_m$ = the modulation angular frequency
$\beta$ = the modulation index
$\phi$ = the rf phase angle.

For the phasor diagram (Fig. 4), let the plane of the figure rotate at frequency $\omega_c$. The phasors representing the filter are stationary and the angle between the signal phasor and one of the filter phasors as a reference is

$$\theta = \phi + \beta\sin\omega_m t.$$

The components of the signal upon the phasors representing the filter response are

$$E\cos(\phi + \beta\sin\omega_m t)$$

and

$$E\sin(\phi + \beta\sin\omega_m t).$$

The time averages are obtained by integration over one cycle of the modulation, giving

$$1/2\pi \int_0^{2\pi} E\cos(\phi + \beta\sin\omega_m t)\,d(\omega_m t)$$

and

$$1/2\pi \int_0^{2\pi} E\sin(\phi + \beta\sin\omega_m t)\,d(\omega_m t).$$

In order to perform the integration, use is made of the identities (found in any book on Bessel functions)

$$\cos(\beta\sin\omega_m t) = J_0(\beta) + 2\sum_{n=1}^{\infty} J_{2n}(\beta)\cos 2\omega_m t$$

and

$$\sin(\beta\sin\omega_m t) = 2\sum_{n=1}^{\infty} J_{2n-1}(\beta)\sin(2n-1)\omega_m t.$$

The time average of either the cosine or sine function vanishes, giving as the results of the integrations

$$EJ_0(\beta)\cos\phi \quad \text{and} \quad EJ_0(\beta)\sin\phi.$$

The vector sum of these components has an amplitude of $EJ_0(\beta)$, which is the correct value as derived by the usual mathematical procedure.

*Example 2:* The amplitude of a spectrum component other than the carrier frequency will now be found.

Let the filter of Example 1 now contain sideband frequency $\omega_c + k\omega_m$ in its pass-band.

If the plane of Fig. 4 is rotating with angular velocity $\omega_c + k\omega_m$, then the phasors representing the filter response are again stationary and the angle $\theta$ between the reference phasor and the FM signal phasor is

$$\theta = k\omega_m t - \phi - \beta\sin\omega_m t,$$

obtained by subtraction of the phase angle of the rotating plane from the phase angle of the FM signal. The components of $e$ upon the two phasors are $E\cos\theta$ and $E\sin\theta$.

The time averages are given by the integrals (expanding the integrands)

$$\mathrm{E}/2\pi \int_0^{2\pi} [\cos k\omega_m t\cos(\phi + \beta\sin\omega_m t) + \sin(k\omega_m t)\sin(\phi + \beta\sin\omega_m t)]\,d(\omega_m t)$$

and

$$E/2\pi \int_0^{2\pi} [\sin k\omega_m t\cos(\phi + \beta\sin\omega_m t) - \cos(k\omega_m t)\sin(\phi + \beta\sin\omega_m t)]\,d(\omega_m t).$$

Using the identities of Example 1 and the orthogonality properties of the trigonometric functions, these time averages reduce to

$$EJ_k(\beta)\cos\phi \quad \text{and} \quad EJ_k(\beta)\sin\phi.$$

The amplitude of the resultant is therefore $EJ_k(\beta)$.

It will now be shown that a filter not containing a spectrum component frequency in its pass-band has a zero average response, even though the filter may pass frequencies within the instantaneous frequency range of the FM signal.

*Example 3:* Let the filter of Example 1 not contain any sideband frequency $\omega_c \pm k\omega_m$. Thus, let the plane of Fig. 4 rotate with angular velocity $\omega_c + \omega_d$ where $\omega_d$ is *not* equal to $\pm k\omega_m$.

The angle $\theta$ is equal to $(\omega_d t - \beta \sin \omega_m t)$ and the components of $e$ upon the reference phasors are $E\cos\theta$ and $E\sin\theta$. The time averages are obtained by integration over a period $T$ of the integrand, giving upon expansion of the integrands:

$$\frac{E}{T}\int_0^T \{\cos\omega_d t\}\{J_0(\beta) + 2\sum J_{2n}(\beta)\cos n\omega_m t\} + \{\sin\omega_d t\}\{2\sum J_{2n-1}(\beta)\sin(2n-1)\omega_m t\}\,dt$$

and

$$\frac{E}{T}\int_0^T \{\sin\omega_d t\}\{J_0(\beta) + 2\sum J_{2n}(\beta)\cos 2\,n\omega_m t\} - \{\cos\omega_d t\}\{2\sum J_{2n-1}(\beta)\sin(2n-1)\omega_m t\}\,dt.$$

Since $\omega_d$ is assumed not equal to any multiple of $\omega_m$, each term of these integrations vanishes by reason of the orthogonality of the trigonometric functions, or the time-averaged response is zero.

The general expression for any spectrum component of an angularly modulated signal can be similarly derived, as is shown in the next example.

*Example 4:* Let the general FM signal be $e = E\cos[\omega_c t + \beta\int_0^{\omega_m t} F(\omega_m t)]d(\omega_m t)$, where $\beta$ is the usual modulation index and $F(\omega_m t)$ is the function describing the way in which the frequency varies.

For simplicity, write $e = E\cos[\omega_c t + g(t)]$ where $g(t)$ describes the angular modulation.

Consider a filter which accepts frequency $\omega_c + \omega_d$ in its pass-band.

Then, just as in Example 3, the angle $\theta$ between $e$ and the reference phasor is $\theta = \omega_d t - g(t)$ and the components of the signal phasor upon the phasors representing the filter response are $E\cos\theta$ and $E\sin\theta$.

The time averages over one period of the angle $\theta$ are:

$$E/T\int_0^T \cos\{\omega_d t - g(t)\}\,dt \quad \text{and} \quad E/T\int_0^T \sin\{\omega_d t - g(t)\}\,dt.$$

Inasmuch as $g(t)$ is a periodic function describing the angular modulation, we may expand $\cos g(t)$ and $\sin g(t)$ into Fourier series giving:

$$\cos g(t) = \sum_{n=0}^{\infty}(b_n\cos n\omega_m t + a_n\sin n\omega_m t)$$

$$\sin g(t) = \sum_{n=0}^{\infty}(d_n\cos n\omega_m t + c_n\sin n\omega_m t).$$

Expanding the integrands and using these Fourier expansions, we note from the orthogonality of the trigonometric functions that the integrals are zero except when $\omega_d = \pm k\omega_m$ with $k = 0, 1, 2\cdots$.

In the latter case, the time averages become

$$E/2(b_k \pm c_k) \quad \text{and} \quad E/2(a_k \mp d_k),$$

whereupon the amplitude of the component of frequency $\omega_c \pm k\omega_m$ is

$$E/2[(b_k \pm c_k)^2 + (a_k \mp d_k)^2]^{1/2}.$$

This result is a perfectly general one and requires only a Fourier analysis of the cosine and sine of the modulation function.

The pair of orthogonal phasors used to represent the filter response may be replaced by a single phasor at any arbitrary angle in the diagrams as shown in Fig. 5. The time average of the component of the FM signal phasor upon this single filter phasor is obtained in the usual fashion, and then is maximized with respect to the arbitrary angle. The correct spectrum amplitudes are obtained in this way verifying the physical concept of the filter response adjusting its phase in such a way as to extract a maximum of energy from the signal. For illustration, Example 1 is repeated by this method. It is required to obtain the carrier amplitude of a sinusoidally modulated FM signal.

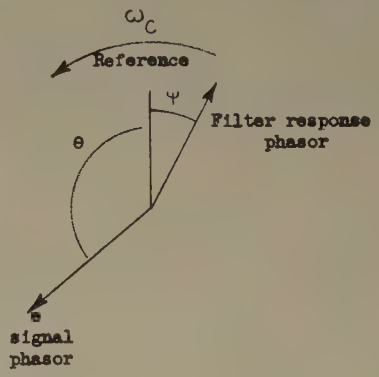


Fig. 5—Representation of filter voltage by a single phasor.

Suppose the phasor representing the filter response is at an angle $\Psi$. Let the FM signal be $e = E\cos(\omega_c t + \phi + \beta\sin\omega_m t)$, and let the plane of the figure be rotating at angular velocity $\omega_c$. Then the component of $e$ upon the filter phasor is $E\cos(\phi + \Psi)$, which gives an integrated time average of $EJ_0(\beta)(\cos\phi\cos\Psi - \sin\phi\sin\Psi)$. This is maximized with respect to $\Psi$ by setting the derivative equal to zero, giving $\tan\Psi = -\tan\phi$, whereupon the amplitude reduces to $EJ_0(\beta)$.

The component view may be extended without difficulty to amplitude-modulated signals, and to signals which are modulated both in amplitude and angle. For example, Fig. 6 shows an amplitude-modulated signal phasor of angular frequency $\omega_c$ and two phasors for the filter response at this frequency. Both the signal phasor and the filter phasors are stationary on the plane rotating at angular velocity $\omega_c$. The components of the AM signal phasor are obtained by projection upon the axes, and the time average gives the usual result.

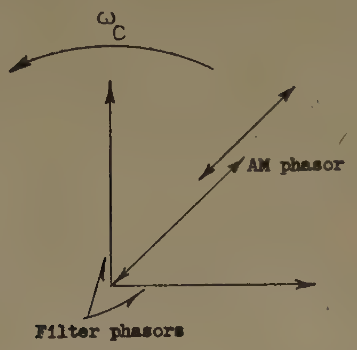
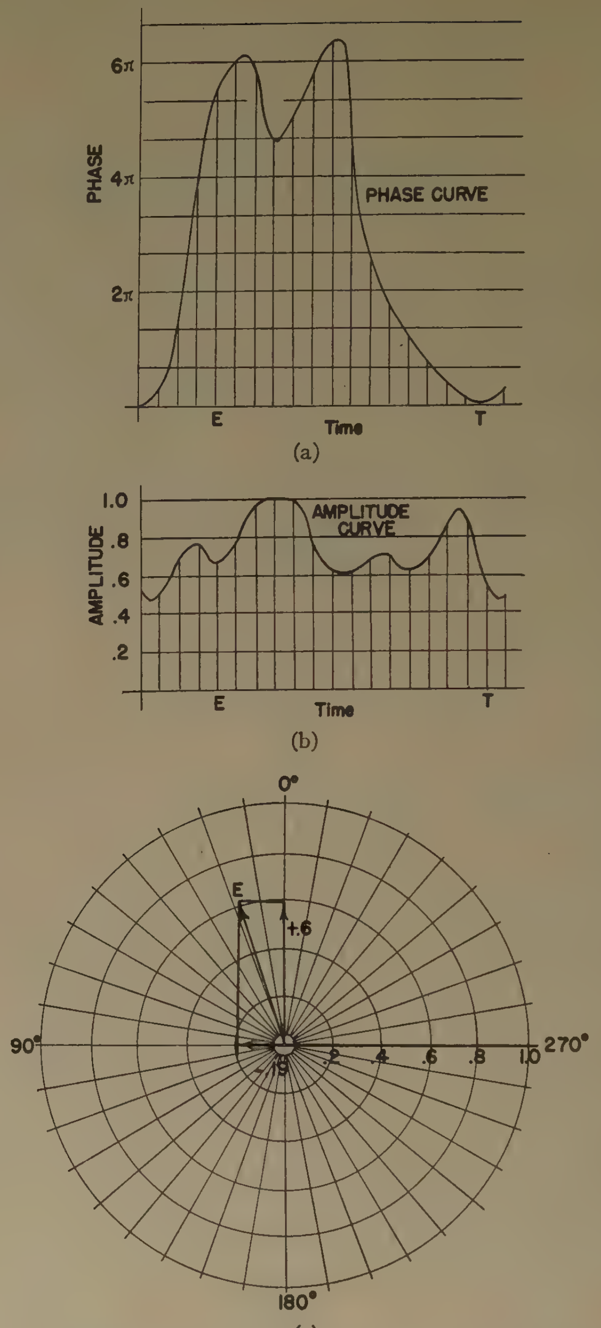


Fig. 6—Representation of an AM signal.

The component method lends itself to graphical analysis of the spectrum of modulated signals, and in fact a good deal of use has been made of this approach in treating certain problems where the modulation waveform has been given in graphical form. The projections of the signal phasor can be measured for any required number of positions of the phasor over a modulation cycle and the results averaged without recourse to integrations or Fourier analysis.

A simple example is presented in Fig. 7, where one seeks the amplitude of the carrier frequency component of a hypothetical signal which is modulated both in phase and amplitude, and where the modulation waveforms for one period are given in graphical form. Fig. 7(a) shows the phase variation of the signal and Fig. 7(b) shows the normalized amplitude variation. (For simplicity the periods have been chosen as equal although generally this will of course not be the case.) The time interval of one period is divided into any convenient number of parts, and for each value of time so chosen the amplitude and phase of the signal are replotted as a phasor, as shown in Fig. 7(c) for the point marked *E*. The components of this phasor upon the axes are easily read and tabulated. The numerical averages of the tabulated components are obtained, squared, added and the square root of the sum of the squares of the averages gives the carrier amplitude.



(a)

(b)

(c)

Fig. 7—Graphical procedure for calculation of the radio spectrum of an amplitude- and frequency-modulated signal.

## CONCLUSION

In the 1922 PROCEEDINGS OF THE I.R.E., Carson[1] first presented the spectrum solution for a sine wave angularly-modulated signal. He then stated that, "The foregoing solutions, though unquestionably mathematically correct, are somewhat difficult to reconcile with our physical intuitions and our physical concepts of such 'variable frequency' mechanisms as for example the siren."

It is just this difficulty which the component theory has attempted to resolve. In brief review, this theory considers the time average of the response of a band selective filter to a modulated signal. The existence of a spectrum component is indicated by the presence of a finite output signal caused by a net energy transfer from the signal to the selective network. This approach yields results identical with those given by the usual mathematical techniques. Its usefulness lies not in its verification of the mathematics, but in furnishing concepts which aid in attacking new problems and which treat all types of modulation from the same viewpoint.

## ACKNOWLEDGMENT

The authors wish to express their appreciation for the many helpful suggestions and criticisms contributed by the Circuits Section of Sylvania Physics Laboratories.

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[1] John R. Carson, "Notes on the theory of modulation," PROC. I.R.E., vol. 22, pp. 57–64; February, 1922.

# Methods of Measuring Adjacent-Band Radiation from Radio Transmitters*

NEAN LUND†, ASSOCIATE, IRE

***Summary***—**A review of three possible methods of measuring or estimating adjacent-band radiation characteristics of a radio transmitter is given. These three methods differ in the type of signal applied to the transmitter and may be termed the two-tone, normal signal, and thermal noise methods. Measurements on a multichannel single-sideband transmitter using each of these methods are presented to show that there is a good correlation between the normal signal and thermal noise methods.**

**An empirical method for calculating the slope of the adjacent-band radiation as a function of frequency from the measured two-tone distortion values is given, and the measured and calculated slopes are shown to be in fairly good agreement.**

## INTRODUCTION

GENERALLY, a radio transmitter has some radiation outside its necessary transmission band, and this radiation may be sufficient to interfere with services occupying other channels. This radiation includes many modulation products generated because of nonlinearity in the transmitter circuits. Usually, the radiation which appears as harmonics of the modulated frequencies can be reduced to satisfactory values by adding selective circuits. Similarly, the very-low-frequency modulation products are sufficiently suppressed by the normal selectivity of the system. However, the radiation which appears just outside the transmitted band cannot be easily reduced by using ordinarily selective circuits.[1] In this paper we will be concerned only with the radiation which falls just outside the transmitted band, and our purpose will be to describe methods of measuring this adjacent-band radiation and to give some results obtained when using these methods on a radio-telephone transmitter.

Three possible methods of measuring or estimating adjacent-band radiation will be described. These are:

1. Measure certain distortion products when two tones of equal amplitude are modulating the transmitter.
2. Measure the radiation in the regions adjacent to the transmitted band when the normal modulating signals are applied to the transmitter.
3. Measure the radiation in the regions adjacent to the transmitted band when thermal noise of equivalent bandwidth and amplitude is substituted for the normal modulating signals.

These three methods of measuring adjacent-band radiation have been applied to a newly designed Western Electric Type LD-T2 single-sideband reduced-carrier transmitter. This transmitter normally operates in the frequency range of 4 to 23 mc and is capable of carrying four telephone channels when used with the proper terminal equipment. The nominal over-all bandwidth of the transmitter is 12 kc consisting of two nominal 6-kc channel groups. These two channel groups are located on either side of the reduced carrier. The reduced-carrier amplitude is normally transmitted 26 db below the rated envelope peak power of 4 kilowatts, and is placed at midband frequency. This transmitter is designed primarily for telephone service, and therefore we are interested in the adjacent-band radiation that results when four talkers are loading the transmitter.

## TWO-TONE DISTORTION MEASUREMENTS

The first or two-tone method for estimating adjacent-band radiation may be carried out by applying two tones of equal amplitude to the transmitter and measuring with a selective analyzer the amplitude of the modulation products which results. A block diagram of the circuit used in making the two-tone tests is shown in Fig. 1. Distortion products of any order of modulation may be measured by using this method. However, in the study of distortion products which lie just outside the transmitted band only odd-order products need be considered.

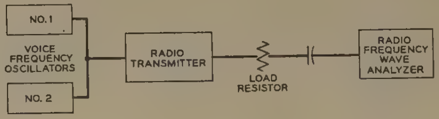

Fig. 1—Block diagram of circuit used in making two-tone distortion measurements.

The results of making the usual two-tone distortion measurements on the LD-T2 transmitter are shown in Fig. 2. The solid curves show the measured ratio of the

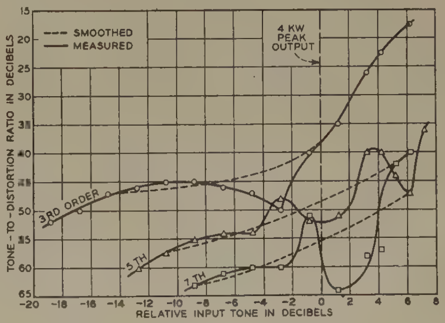

Fig. 2—Tone-to-distortion ratios plotted against input tone amplitude for the LD-T2 radio transmitter. The solid curves show measured values and the dotted curves show smoothed values.

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* Decimal classification: R254.1×R148.13. Original manuscript received by the Institute, June 2, 1950; revised manuscript received, September 15, 1950.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] I. J. Karr, "Some notes on adjacent channel interference," PROC. I.R.E., vol. 22, pp. 295–313; March, 1934.

power in one of the two output tones to the power in a single distortion product as a function of the transmitter input tone amplitude. Data are plotted for third-, fifth-, and seventh-order products. When the transmitter is operating at its envelope peak-power rating of 4 kw, these curves show that the third-order distortion ratio is 38 db, the fifth-order distortion is 52 db, and the seventh-order distortion is 59 db. Some very general conclusions might be drawn from these two-tone curves. Since the distortion decreases with the order of modulation product, it might be expected that the adjacent-band radiation would decrease with frequency from the center of the transmitted band. Also, since the average amplitude of the distortion decreases as the signal input decreases, it would be expected that the adjacent-band radiation would decrease as the signal input is decreased. These conclusions are qualitative and of small value in determining the actual amplitude of adjacent-band radiation. However, such distortion measurements on one transmitter can be compared to similar measurements on another transmitter to obtain a reasonably accurate idea of the relative adjacent-band radiation performance.

### Measurements Made with Normal Signal Loading

In order to carry out the second or normal signal method it was necessary to provide four sources of speech signals which could be used to modulate all four channels. Also, in order to be able to use these four speech sources it was necessary to have channel shifter equipment which allowed two 2.5-kc speech channels to be placed in the channel group frequency range from 100 to 6,000 cycles per second. A block diagram of the circuit used in the normal signal loading tests is shown in Fig. 3. Since the transmitter is to be used in two-way telephone communication, a subscriber would be expected to be talking on the circuit about 50 per cent of the time his call is in progress. Therefore, to represent this condition, recordings were made of conversational type speech, that is, the speech recorded was one side of a two-way conversation. Each record was made from a different conversation. These records were made on magnetic tape recorders which were capable of continuously playing a one-minute recording.

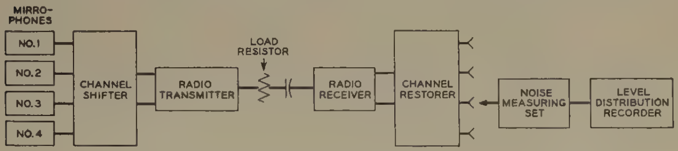

Fig. 3—Block diagram of circuit used in making the normal signal loading tests.

The adjacent-band radiation was measured with the Western Electric type LD-R1 single-sideband receiver, tuned to definite spot frequencies outside the band of the transmitter. Measurements were made at 3-kc intervals out to 40 kc from the transmitted carrier using a receiving bandwidth of 2.5 kc.

In order to actually measure fluctuating adjacent-band radiation, it is necessary to use some instrument which is capable of giving readings which can be interpreted. This is particularly important when measuring widely fluctuating modulation noise amplitudes such as are produced by speech modulation. An instrument which is quite satisfactory for this purpose is a Level Distribution Recorder.

The results of these adjacent-band radiation measurements have been plotted as ratios of the speech amplitude in one channel to the adjacent-band radiation amplitude in a band of the same width. These results are shown by the dashed curves plotted in Fig. 4 for three different input volumes which correspond to normal input, 5 vu above and 5 vu below normal. These three loading conditions give transmitter output volumes of 50, 55, and 59.5 vu per channel.

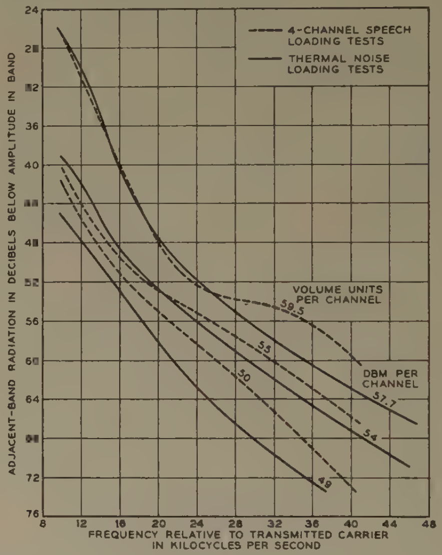

Fig. 4—Comparison of results of four-channel speech loading tests and thermal noise loading tests on the LD-T2 radio transmitter.

### Measurements Made with Thermal Noise Loading

In the third or thermal noise method, it is necessary to have a source of thermal noise which has about the same band of frequencies as the speech band. This is necessary in order that the intermodulation between the noise frequencies will produce the same range of adjacent-band components as are produced by the speech modulation. Since the LD-T2 transmitter has two 5.9-kc channel groups, it was necessary to have two independent sources of thermal noise each producing a band from 100 to 6,000 cycles. Another requirement is that the noise energy per unit of bandwidth shall be constant with frequency. A convenient source of such thermal noise was the 5.9-kc channel group outputs of an LD-R1 single-sideband receiver. A block diagram of the circuit used in the thermal noise tests is shown in Fig. 5.

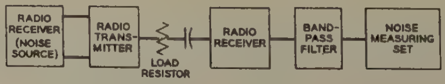

Fig. 5—Block diagram of circuit used in making the thermal noise loading tests.

The adjacent-band noise was measured by using a Western Electric type 2B Noise Measuring Set. There was very little fluctuation in the reading so that simply recording the reading of this instrument was sufficient. Filters were used in the analyzer receiver output to limit the received bandwidth to 2.5 kc. Measurements were made with the analyzer receiver tuned to the transmitted band so as to obtain the amplitude of the transmitted thermal noise in a 2.5-kc band. The measured adjacent-band modulation noise was then compared to this amplitude in order to obtain a ratio of single channel power to adjacent-band radiation power. The results are shown by the solid curves in Fig. 4. Measurements were made in 3-kc steps out to about 40 kc from the carrier for three values of noise input to the transmitter. As will be explained later, these three loading conditions approximate normal loading and conditions 5 db above and 5 db below normal. The adjacent-band radiation for about normal loading is shown by the curve for 54 dbm power output per channel.

## CORRELATION OF ADJACENT-BAND RADIATION MEASUREMENTS AND TWO-TONE DISTORTION MEASUREMENTS

Since the adjacent-band radiation measurements require a fairly elaborate testing arrangement, it would be desirable to be able to predict the adjacent-band radiation for any transmitter on the basis of two-tone tests. Some calculations have been made for the LD-T2 transmitter to determine a correlation between two-tone distortion and adjacent-band radiation. These calculations have been made using the following assumptions:

1. It is assumed that a uniform 12-kc band of noise is applied to the transmitter amplifier and that this noise can be analyzed as being equal to a large number of equally spaced sine wave signals having equal amplitudes and random phase in the 12-kc band.
2. The intermodulation between these sine wave signals gives a distribution of modulation products which can be determined for each order of modulation product.
3. The contribution of each distribution to the total distribution is found by multiplying each distribution by the two-tone distortion ratio for that particular order of modulation product.
4. The two-tone distortion ratios used in step 3 are the values measured for an envelope peak tone input which is 6 db greater than the total average noise power input.
5. The modulation noise at any frequency can be found by summing up the contributions found in step 3 for each order of modulation product.
6. The addition of the distributions in step 5 is made on the basis of power addition of the individual distributions.

In order to carry out the procedure which has been outlined, it is necessary to determine the frequency distribution of modulation products for each order of modulation product. A method for determining a distribution of modulation products is to assume a definite number of modulating tones and calculate the number of modulation products which fall into a bandwidth equal to the spacing of the modulating tones. Such a process has been carried out by W. R. Young of the Bell Telephone Laboratories so as to determine formulas for third-order modulation products. A distribution of third-order products, calculated from third-order formulas, is shown in Fig. 6. The curve plotted in this figure shows the relative number of modulation products as a function of the relative frequency from the center of the modulating band. It has been assumed that this curve can be used to determine the distributions for higher orders of modulation, as well as for third-order, by stretching the abscissa values in proportion to the order of modulation. This is a simplifying assumption and no estimate of the error that results is given here.

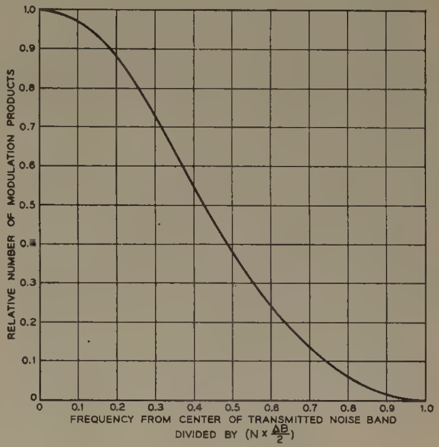

Fig. 6—Distribution of modulation products used to calculate adjacent-band radiation. *N* is the order of modulation product and *B* is the width of the transmitted noise band.

It is desired, then, to multiply the frequency distributions by the corresponding two-tone distortion ratio shown in Fig. 1. The data actually used in this figure were the smoothed or dashed curves shown. The reason for using the smoothed curves is to minimize the effects of the dips or cancellation points in the two-tone curves. The cancellation effect takes place over a small range of amplitudes which can be observed on a two-tone test. However, for noise inputs the amplitudes have a large range of values and the cancellation effect takes place over only a small portion of this range. For noise amplitudes slightly greater or less than the critical amplitude, the modulation products are much stronger so that products generated when the noise envelope is at the critical amplitude contribute very little to the total distortion.

The results of both measured and calculated adjacent-band radiation are shown in Fig. 7, and show that the calculated values are in fairly good agreement with the measured values for all three values of loading. It may be noted that the procedure for calculating the adjacent-band radiation has been developed through the use of empirical methods, and therefore it would be desirable to obtain some further checks by application of the method to other types of transmitters.

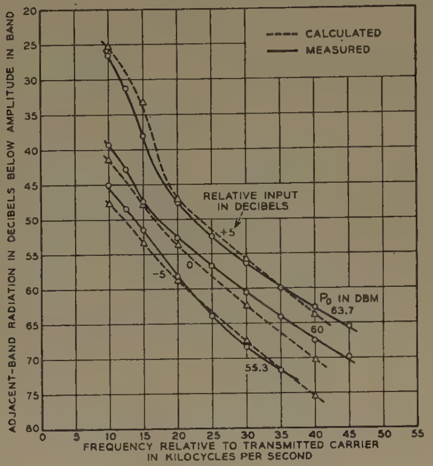

Fig. 7—Comparison of measured and calculated adjacent-band radiation for the LD-T2 radio transmitter. The solid curves show measured values for the thermal noise loading tests. The dashed curves show calculated values.

## COMPARISON OF NORMAL SIGNAL AND THERMAL NOISE TESTS

A direct comparison of the measured results obtained by the speech and thermal noise methods is shown in Fig. 4. The comparison data show that in general the slopes of the adjacent-band radiation for both tests are the same, and that the absolute values of the adjacent-band radiation for both tests are about the same. In Fig. 4 the three curves for speech loading are shown in terms of vu output per channel and the three curves for thermal noise loading are shown in terms of dbm output per channel. These results indicate that on the average a speech loading which is 1 vu greater than the noise loading in dbm will produce about the same amount of adjacent-band radiation.

## CONCLUSIONS

It is concluded that the thermal noise method of making measurements gives results which are essentially the same as those obtained with normal signal loading. The thermal noise method is much superior to the normal signal method in that it requires considerably less equipment for generating the modulating signals and for making the measurements. Also in the thermal noise method there is very little fluctuation in the amplitudes to be measured, and therefore the readings may be obtained quite rapidly and accurately.

## ACKNOWLEDGMENT

The author wishes to express his indebtedness to the Overseas Radio Group at Deal, N. J., for their assistance in making the measurements, and to F. B. Llewellyn for his valuable suggestions in connection with the thermal noise method.

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# A Very-High-Frequency—Ultra-High-Frequency Tail-Cap Antenna*

LOUIS E. RABURN†, SENIOR MEMBER, IRE

***Summary*—A description is given of a broad-band vhf-uhf antenna for a small, single-engine type of aircraft consisting of the vertical stabilizer that is tip-fed by the metal tail cap. The resulting antenna covers the frequency range of 100 to 400 mc, is vertically polarized, and has radiation patterns that are essentially omnidirectional. It is a zero-drag type of antenna that is very adaptable to aircraft production techniques.**

**The tail cap, together with the rest of the vertical stabilizer, performs as a vertical sleeve-fed stub on an unsymmetrical ground plane. For this reason, the antenna radiation fields are not perfectly omnidirectional, but they are adequate for two-way communication in the frequency range 100 to 400 mc, and in this range the standing wave ratio on RG-8/U 50-ohm coaxial cable is less than 2 to 1.**

**The predicted radiation field-strength patterns for a specified transmitter power are calculated from measured radiation field patterns of the scale-model aircraft and full-scale standing-wave ratio measurements.**

## INTRODUCTION

THE POSSIBILITIES of exciting aircraft wings, fuselage, and empennage for use as low-drag or zero-drag antennas at frequencies above the medium frequency band has been investigated over a period of years by a number of people, both in this country and abroad.[1-6] An extensive study of tip-

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* Decimal classification: R326.21×R326.7. Original manuscript received by the Institute, May 29, 1950; revised manuscript received, September 5, 1950.

† Electronics Research, Inc., Evansville, Ind.

[1] M. P. Hanson, "HF Collecting and Radiating Structure," U. S. Patent No. 2,044,779; June 23, 1936.

[2] S. Zisler, "Theory and Technique of Antennas, Part I," pp. 24–30. This is a German document which has been translated and declassified. It is available at the Central Air Documents Office, Wright Field, Ohio.

[3] W. A. Johnson, "Recent developments of aircraft communication aerials," *Jour. IEE*, Part IIIA, p. 452; 1947.

[4] F. H. Behrens, "Drag cut with plastic antenna housings," summarized by McLarren, *Aviation Week*, vol. 49, pp. 18–24; November 29, 1948.

[5] G. E. Beck, "Suppressed aircraft aerials," *Wireless World*, vol. 26, p. 468; December, 1949.

[6] J. V. N. Granger, "Shunt-excited flat-plate antennas with applications to aircraft structures," PROC. I.R.E., vol. 38, p. 280; March, 1950.

excited wings and stabilizers was sponsored several years ago by the Army Air Force and was performed at the Rocky Point Laboratories of RCA.[7]

This study showed that it was possible to excite a wing by insulating and feeding the outer one-fifth of its length. The resulting antenna bears some physical resemblance to a sleeve stub and subsequent work has confirmed the fact that most of the electrical characteristics are also similar.[8] It has been found that sleeve-type stubs with large-diameter elements of optimum proportions have very desirable impedance characteristics over broad frequency bands.[9]

When it was required to develop a zero-drag vertically polarized omnidirectional antenna for use with vhf and uhf communications equipment in a small, single-engine aircraft of relatively conventional type, the tip-fed vertical stabilizer configuration seemed most feasible. The development of this antenna will be described, and because the feeding tip is the vertical stabilizer cap, the antenna will be referred to as the tail-cap antenna throughout the rest of this paper.

## RADIATION PATTERN STUDY

A tail-cap antenna in the vertical stabilizer of a conventional-type aircraft has similar electrical characteristics to a sleeve stub near the side of a horizontal ground plane. For this reason, it isn't considered possible to change the radiation patterns very much by changing the shape or size of the tail-cap alone. It is, therefore, desirable to perform a preliminary pattern study using scale-model techniques to determine whether or not the radiation characteristics of the tip-excited tail antenna would be acceptable before the full-scale prototype antenna is developed.

This preliminary pattern study was performed using the measuring range at the roof laboratory shown in Fig. 1. A one-eighth scale, precision model of the aircraft was made of wood and sprayed with copper to provide a highly conductive coating. A scale model of the capacity-hat tail-cap antenna with the proposed feeding and matching configuration was attached to the vertical stabilizer of the model in Fig. 1. The model antenna had an SWR of less than 4 to 1 throughout the model frequency range of 800 to 3,200 mc, and did not need any special matching system for the pattern measurements.

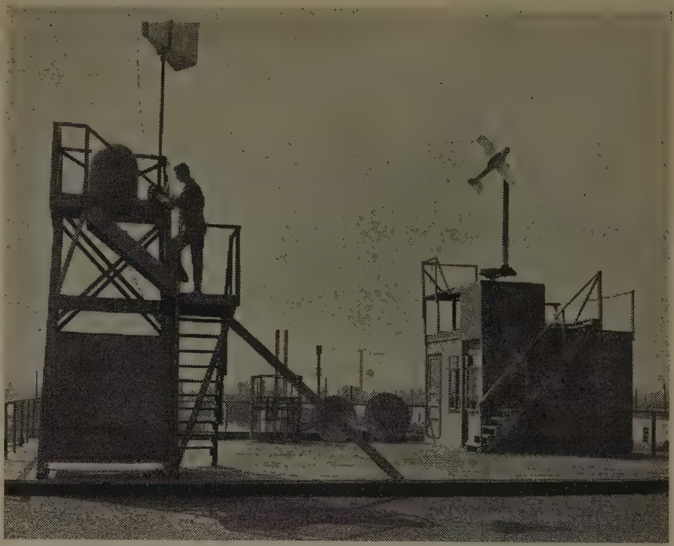
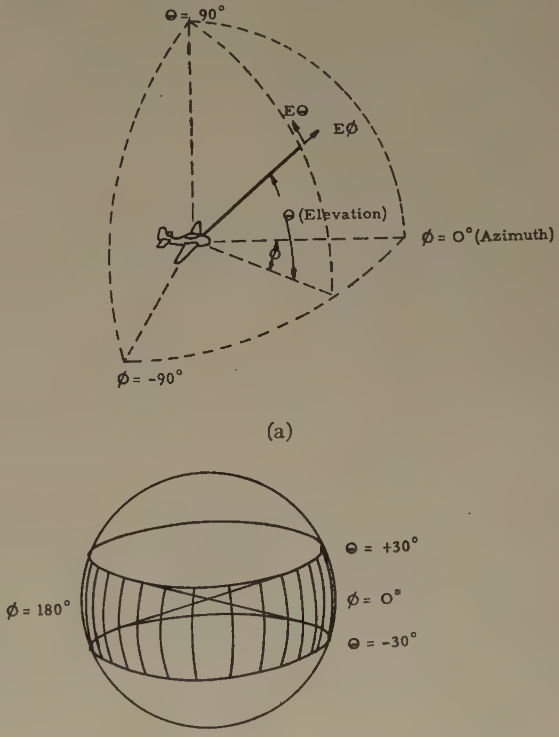


Fig. 1—Scale-model aircraft and radiation pattern-measuring range.



(a)

(b)

Fig. 2—(a) Radiation pattern co-ordinate system. (b) Equatorial belts for region $\theta = +30°$ to $-30°$.

A complete set of principal plane patterns and cone-angle patterns was recorded for both $E_\theta$ (level-flight vertical polarization), and $E_\phi$ (level-flight horizontal polarization) at simulated frequencies of 130, 250, 300, and 362 mc. The $E_\theta$ and $E_\phi$ polarizations are illustrated in the aircraft co-ordinate system of Fig. 2(a). The set of principal-plane patterns for 250 mc is shown in Fig. 3 to best indicate the general shape of the patterns for both $E_\theta$ and $E_\phi$ polarization. The cone-angle patterns, plotted on a basis of relative directivity only, were then integrated and computations were made[10] to convert the relative patterns into field-strength contour patterns at a distance of one mile from the aircraft for a transmitting power of one watt.

The complete set of patterns is very bulky, and the important data within the cone angles of +30° and −30° can be consolidated in the more convenient form of equatorial belt diagrams of the type shown in Figs. 2(b) and 4. In these diagrams a rectangular co-ordinate system is used with azimuth plotted as abscissas and cone angle as ordinates. Contours of equal field strength are formed on the diagram by joining points of constant field strength obtained from the complete set of radiation patterns for each test frequency. The radiation patterns of the tail cap are practically symmetrical about the axis of the aircraft, and therefore, only the port half of each spherical radiation pattern (for vertical polarization) is shown. Those areas in the equitorial belt where the field strength is less than one-half millivolt per meter at one mile for one-watt input are indicated in solid black and represent the prominent nulls which might impair successful communication. Nulls are therefore immediately obvious from an inspection of the equatorial belt diagrams, and as can be seen there are only a few null areas, and these are so small they would not seriously affect the communications during flight.

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[7] P. S. Carter, H. C. Lawrence, R. S. Wehner, and J. D. Woodward, "HF communications development program, Item 1," RCA Report, pp. 7–48; June, 1947.

[8] E. L. Bock, J. A. Nelson, and A. Dorne, "Very High Frequency Techniques," pp. 119–123, McGraw-Hill Book Co., New York, N. Y.; June, 1947.

[9] N. E. Lindenblad, "Television transmitting antenna for Empire State Bldg.," *RCA Rev.*, vol. 3, p. 387; April, 1939.

[10] F. E. Terman, "Radio Engineer's Handbook," McGraw-Hill Book Co., New York, N. Y., pp. 782–784; 1943.

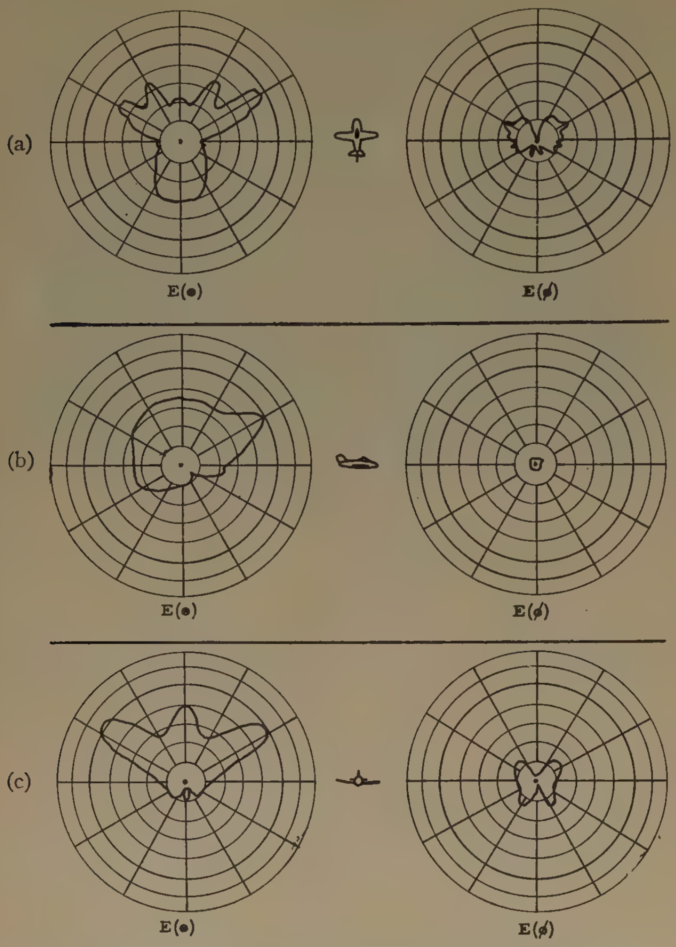

Fig. 3—Principal-plane patterns at 250 mc. (a) Horizontal plane. (b) Fore-and-aft plane. (c) Transverse plane.

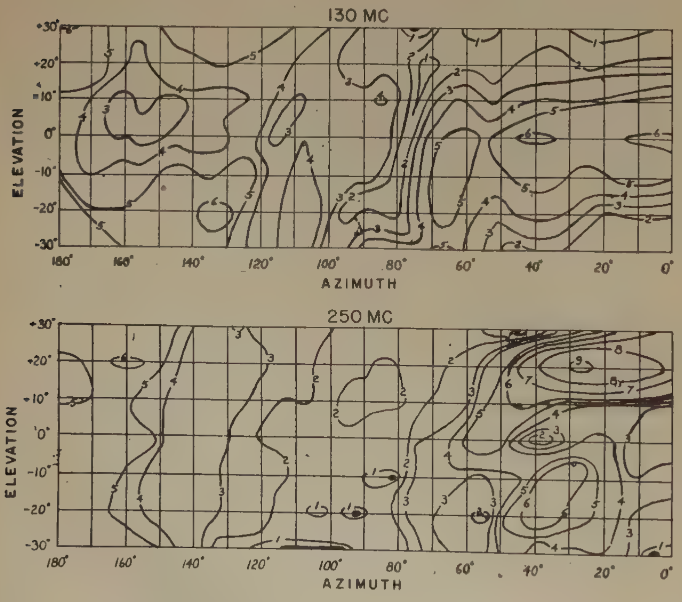

Fig. 4—Absolute field strength in equatorial belt. (Field-strength units—millivolts per meter for one watt at a distance of one mile.) Black areas represent field strength less than 0.5 millivolt per meter for one watt at a distance of one mile. Polarization—$E_\theta$ vertical polarization when aircraft is in level flight.

The ultimate purpose of the absolute field-strength patterns is to predict what the maximum ranges for acceptable operation will be when the antenna is used with specified radio equipments. These predictions are based on operational characteristics of the pertinent equipment, propagation data such as given in the literature,[11] assumed cable attenuation, and so forth. The predicted ranges depend upon the altitude of the aircraft, and are usually plotted as curves of maximum range versus altitude. The predictions from this model study indicate that the tail-cap antenna has satisfactory directivity characteristics, and it was concluded that a properly matched tail-cap antenna would give acceptable results when installed on the actual aircraft.

## Development of Electrical Prototype

After the radiation pattern study was completed with acceptable results, the development of the full-scale electrical prototype tail-cap was begun. This phase was performed using a full-scale mock-up of the vertical portion of the tail above the horizontal stabilizer. The mock-up was constructed of plywood covered with copper and hardware cloth as shown in Fig. 5.

The mock-up was set on a 15-foot-square ground plane to simulate the empennage and fuselage for purposes of VSWR and matching measurements. An eight-foot length of RG-8/U coaxial cable was connected between the antenna and a precision slotted line located in the hut below.

The tail cap of the aircraft is 16 inches tall and provides the best solution for meeting all the design requirements. The all-metal stabilizer tip is isolated from the remainder of the stabilizer by a small strip of dielectric material, thus eliminating the need for a fiberglas stabilizer cap made by special production techniques.

The copper prototype tail cap was mounted on the stabilizer rib by two fiberglas pillars as shown in Fig. 5. The tail-cap face of the feeding gap connects to the center conductor of a coaxial cable receptacle, and the stabilizer face of the gap connects to the outer shield of the coaxial receptacle by a metal pyramid having a base $4\frac{1}{2}$ inches by 9 inches across. The pyramid provides a smooth transition between the coaxial receptacle and the feeding gap, and the slope of the pyramid was chosen for best broad-band performance.

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[11] C. R. Burrows, et al., "Radio Wave Propagation," Academic Press, New York, N. Y.; 1949.

The impedance-matching problem was solved in part by the use of an optimum pyramid configuration, but in order to meet the SWR requirements at the low end of the frequency band, it was found necessary to employ a shunt inductance across the gap. The final configuration of the shunt consists of two copper wires 0.125-inch diameter and $3\frac{1}{2}$ inches long placed in the gap near the leading edge of the stabilizer.

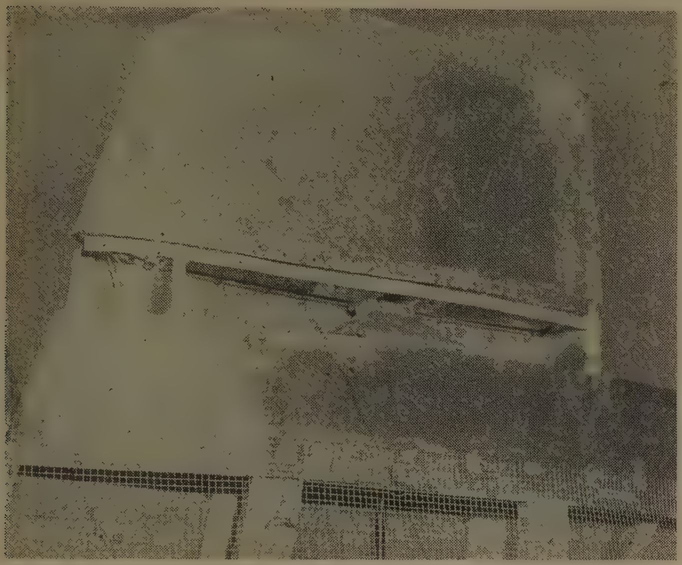

Fig. 5—Full-scale mock-up, prototype antenna.

Fig. 6 is a plot both of the SWR of the antenna as measured with 22 feet of RG-8/U cable as lead-in and also with the SWR at the antenna terminal corrected for the attenuation of the cable. The resultant SWR in both cases is less than 2 to 1. All final impedance and SWR measurements were made with a fiberglas rudder cap in place and with a fiberglas belt around the capacity hat.

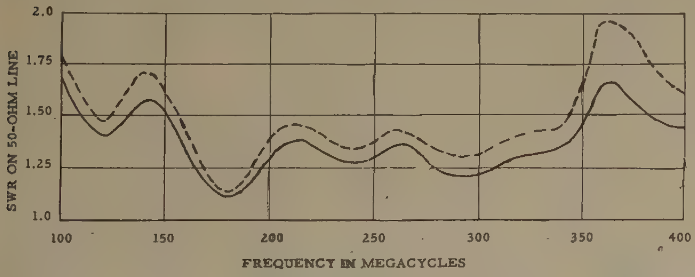

Fig. 6—Standing-wave ratio for tail-cap antenna. (———SWR with 22 feet RG-8/U. - - - - - SWR calculated at antenna.)

## Mechanical Design and Assembly

The mechanical design of the antenna incorporates a $\frac{1}{8}$-inch thick fiberglas belt 3 inches wide and two fiberglas structural spars $4\frac{1}{2}$ inches wide by $\frac{1}{2}$ inch thick shown in Fig. 5. The spars provide structural strength in addition to that of the fiberglas belt around the gap. The section of the rudder tip above the feeding gap in the stabilizer is made of fiberglas to avoid shielding and shunting effects on the tail cap. All fiberglas used is Formica FF-55 which has a dielectric constant of about 7. In order to control the current distribution in the metal part of the rudder, flexible bonding straps are run between the rudder and vertical stabilizer.

For convenience in production assembly, the pyramid and shunt wires are designed so that they may be installed after the tip has been mounted, or, if preferred, the pyramid may be assembled to the tip before the tip is mounted. The top of the pyramid is a small box that can be adjusted to change the over-all height of the pyramid a sufficient amount to compensate for production variations in the width of the feeding gap.

## Conclusions

The tail-cap antenna described in this paper has acceptable electrical characteristics; is easy to construct, install, and service; does not cause any increase in aerodynamic drag; and last, but not least, it does not cause an increase in the weight in the tail of the aircraft. For these reasons, it is considered to be a good antenna for this application.

There are two shortcomings that this type of antenna may have. Some aircraft fuselage types may be so designed as to cause poor radiation patterns and also narrow-band characteristics. In these cases, the performance can not be made acceptable regardless of the tail cap configuration. The second shortcoming is that when this type of antenna is installed on a high-speed aircraft, moisture and ice particles errode the leading edge of the fiberglas belt and destroy its smooth contour. Perhaps this problem, which also exists with all types of fiberglas radomes and covers, will be solved to a great degree by continuing development of plastic materials.

## Acknowledgments

Much credit is due to Richard Anderson and William Barrick of Electronics Research, Inc., for the measurements and development given in this paper, and to Eugene H. Barnard also of Electronics Research, Inc., who contributed considerably to the final configuration of the electrical prototype.

# The Patterns of Antennas Located Near Cylinders of Elliptical Cross Section*

GEORGE SINCLAIR†, SENIOR MEMBER, IRE

*Summary*—**The patterns of small dipole and loop antennas mounted on or near a perfectly conducting cylinder of elliptical cross section are calculated by determining the open-circuit terminal voltages of the antennas when receiving plane waves. The antennas are assumed to be such that they cause negligible distortion of the field when they are open-circuited, thus reducing the problem to that of calculating the diffraction of a plane wave around the cylinder. The results have applications in the design of aircraft antennas and slotted-cylinder antennas.**

## INTRODUCTION

IN THE DESIGN of antenna systems the problem of determining the effect of a nearby obstacle on the performance of the system frequently arises. An antenna may be mounted on or near a metallic structure, such as a building or an aircraft, with resultant distortion of the pattern. In some cases this distortion is undesirable and must be minimized while in others the distortion may be utilized to obtain a more desirable pattern. In either case, a knowledge of the distortion is required in carrying out the design of the system.

An antenna system in which the mounting structure plays an important part is that of an airborne antenna since the aircraft structure itself is often an essential portion of the radiating system. The effects of the aircraft structure on the pattern may be calculated approximately by replacing the complicated shape of the aircraft by a simpler geometrical shape more amenable to calculation. For example, if the antenna is mounted on the fuselage, it is known that an infinitely long circular cylinder is often a good approximation. When the antenna is mounted on a wing or on the empennage, a cylinder of elliptical cross section forms an adequate approximation in most cases.

An analysis of the effect of a nearby structure on the pattern of an antenna has applications also in the design of slotted-cylinder antennas since a slot antenna can be analyzed in terms of an equivalent distribution of magnetic dipoles. A slot antenna located on the surface of a cylindrical metal structure has a directional pattern, and a knowledge of the directive properties is usually required in designing the system. Since the pattern is dependent to some extent on the cross-sectional shape of the cylindrical surface, it is possible to obtain some control of the pattern by properly shaping the cylinder.

The patterns of simple antennas mounted on or near cylinders of circular cross section have been investigated by P. S. Carter[1,2] by a method based on diffraction theory and the reciprocity theorem. In the following, the analysis is extended to antennas located near cylinders of elliptical cross section. The method yields information only on the distant field radiated by the antenna, and hence produces no information on the reactive component of the terminal impedance.

## RELATIONSHIP BETWEEN ANTENNA PATTERNS AND DIFFRACTION THEORY

The application of diffraction theory to the calculation of antenna patterns follows naturally from a consideration of the fields existing around an antenna receiving plane waves from a distant transmitter. Consider, for example, a dipole antenna whose dimensions are small compared with the wavelength and which is located near a given object of arbitrary shape, receiving waves from a given direction. When the antenna is open-circuited at its terminals, it causes negligible distortion of the field which would exist with the antenna absent. The dipole is merely a small probe which samples the component of the electric field parallel to its axis. If this component of the field can be calculated by diffraction theory, for different directions of incidence of the plane wave, the pattern of the dipole as a receiving antenna can be obtained. The distant field radiated by the antenna when transmitting is then determined from the reciprocity theorem. Similarly, if the antenna is a loop whose dimensions are small compared with the wavelength and with its terminals open-circuited, the receiving pattern can be determined from a knowledge of the component of the magnetic field intensity parallel to the axis of the loop.

The diffraction of plane waves by an elliptic cylinder has been considered by several investigators,[3,4] but their solutions are not sufficiently general since they considered only waves incident in directions normal to the axis of the cylinder. In the Appendix, expressions are obtained for the field produced near an infinitely long, perfectly conducting elliptic cylinder when illuminated by a plane wave of arbitrary polarization and traveling in an arbitrary direction.

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* Decimal classification: R120×R326.21. Original manuscript received by the Institute, June 16, 1949; revised manuscript received, August 7, 1950.

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† Formerly, The Ohio State University Research Foundation, Columbus, Ohio; now, University of Toronto, Toronto, Ont., Canada.

[1] P. S. Carter, "Antenna arrays around cylinders," PROC. I.R.E., vol. 31, pp. 671–693; December, 1943.

[2] P. S. Carter, "Antennas and Cylindrical Fuselage," Report No. 895-11, RCA Laboratories, Rocky Point, N. Y.; December 24, 1943.

[3] P. M. Morse and P. J. Rubenstein, "The diffraction of waves by ribbons and slits," *Phys. Rev.*, vol. 54, pp. 895–898; December, 1938.

[4] B. Seiger, "Die beugung einer ebenen elektrischen wellen an einem schirm von elliptischen querschnitt," *Ann. der Phys.*, vol. 27, pp. 626–664; November, 1908.

## PATTERNS OF ANTENNAS NEAR AN ELLIPTIC CYLINDER

In specifying the location of the antenna near the elliptic cylinder, it will be convenient to employ elliptic cylinder co-ordinates $(u, v, z)$ which may be defined by the following transformation from Cartesian co-ordinates:

$$x = d \cosh u \cos v \tag{1}$$

$$y = d \sinh u \sin v \tag{2}$$

$$z = z, \tag{3}$$

where $2d$ is the distance between the foci of the ellipses. The foci are located at the points $(x = \pm d, y = 0)$, as shown in Fig. 1. The surfaces $u$ = constant represent a family of confocal cylinders of elliptical cross section. The semimajor and semiminor axes of an ellipse $u = u_0$ are given by

$$a' = d \cosh u_0 \tag{4}$$

$$b' = d \sinh u_0. \tag{5}$$

An open-circuited antenna of infinitesimal dimensions (a dipole or a loop antenna) is situated at the point $P_1(u_1, v_1, 0)$ outside (or on) the conducting surface of the cylinder $u = u_0$. No loss in generality results from assuming $z_1 = 0$ since the cylinder extends to infinity in both directions along the $z$ axis from $P_1$. The pattern of the antenna is to be calculated by determining the open-circuit terminal voltage as a function of the direction of incidence of the plane wave.

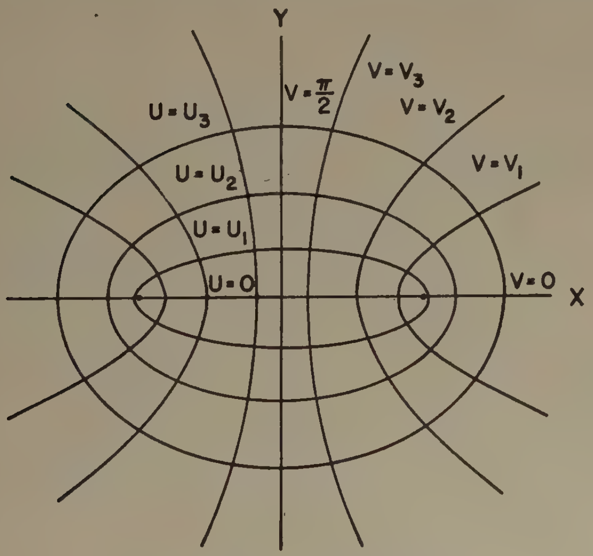

Fig. 1—The elliptic co-ordinate system.

When the antenna is being used for transmission, it produces a distant field which is essentially a spherical wave, so that in specifying the pattern of the antenna it is more suitable to employ a spherical co-ordinate system. The distant field can be resolved into the two spherical components $E_\theta$ (theta-polarized component)[5] and $E_\phi$ (phi-polarized component),[5] as shown in Fig. 2. By the reciprocity theorem[6] it can be shown that the relative pattern for either component can be obtained by determining the open-circuit terminal voltage of the antenna when receiving a plane wave polarized parallel to the component under consideration, the direction of propagation of the plane wave being varied to obtain the pattern.

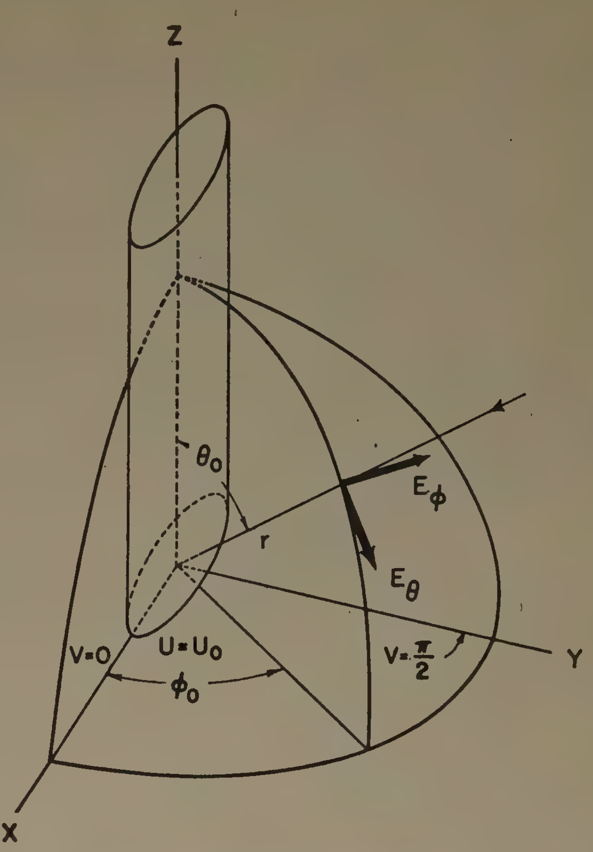

Fig. 2—Co-ordinate system, showing the resolution of the field into components.

While it is possible to calculate the patterns of loop and dipole antennas of any arbitrary orientation with respect to the cylinder, only the cases of dipole and loop antennas whose axes are tangent to co-ordinate curves will be considered. Consider an infinitesimal dipole of effective length $l$, receiving a plane linearly polarized wave, polarized in the direction for measuring the $E_\theta$ component of the transmitting pattern. Let the open-circuit voltage produced be $V_\theta$. This voltage is equal to the product of the effective length and the component of the electric field intensity parallel to the axis of the antenna. Similarly, for a phi-polarized wave, the open-circuit voltage is the effective length times the component of electric intensity parallel to the axis. For a loop antenna of area $S$, the open-circuit voltage is equal to the product of its effective length $l = j\omega\mu S$ and the component of the magnetic intensity parallel to the axis of the loop. Using the expressions for the field components derived in the Appendix, the following are the expressions for calculating the open-circuit terminal voltages for the different cases:

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[5] I.R.E. Standards on Antennas, Definitions of Terms, 1948.

[6] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., p. 476; 1943.

*Radial Dipole*

$$V_\theta = \frac{l}{h_1} \left. \frac{\partial^2 \Pi_z}{\partial u \partial z} \right|_{P_1} \tag{6}$$

$$V_\phi = -j \frac{\omega \mu l}{h_1} \left. \frac{\partial \Pi_z^*}{\partial v} \right|_{P_1}, \tag{7}$$

where $\Pi_z$ and $\Pi_z^*$ are the Hertzian potentials[7] used in calculating the field (see Appendix).

*Circumferential Dipole*

$$V_\theta = \frac{l}{h_1} \left. \frac{\partial^2 \Pi_z}{\partial v \partial z} \right|_{P_1} \tag{8}$$

$$V_\phi = j \frac{\omega \mu l}{h_1} \left. \frac{\partial \Pi_z^*}{\partial u} \right|_{P_1}. \tag{9}$$

*Dipole Parallel to the Axis*

$$V_\theta = l k^2 \sin^2 \theta_0 \Pi_z \big|_{P_1} \tag{10}$$

$$V_\phi = 0. \tag{11}$$

*Loop with Axis Radial*

$$V_\theta = -\frac{k^2 S}{h_1} \left. \frac{\partial \Pi_z}{\partial v} \right|_{P_1} \tag{12}$$

$$V_\phi = j \frac{\omega \mu S}{h_1} \left. \frac{\partial^2 \Pi_z^*}{\partial u \partial z} \right|_{P_1}. \tag{13}$$

*Loop with Axis Circumferential*

$$V_\theta = \frac{k^2 S}{h_1} \left. \frac{\partial \Pi_z}{\partial u} \right|_{P_1} \tag{14}$$

$$V_\phi = j \frac{\omega \mu S}{h_1} \left. \frac{\partial^2 \Pi_z^*}{\partial v \partial z} \right|_{P_1}. \tag{15}$$

*Loop with Axis Parallel to Axis of Cylinder*

$$V_\theta = 0 \tag{16}$$

$$V_\phi = j \omega \mu k^2 S \sin^2 \theta_0 \Pi_z^* \big|_{P_1}. \tag{17}$$

In these expressions, $\theta_0$ defines the direction of incidence with respect to the axis of the cylinder (see Fig. 2), $h_1$ is the metrical coefficient, and $k^2 = \omega^2 \mu \epsilon$.

In each of the equations (6) through (17), the derivatives are to be evaluated at the point $P_1$ where the antenna is located. The total terminal voltage when waves of both theta and phi polarizations are present is the sum of $V_\theta$ and $V_\phi$, taking due account of phase differences between the components of the incident wave.

The effective length of the antenna is easily obtained from the expressions for the received voltages. The vector effective length $\boldsymbol{h}$ of the antenna[8] can be resolved into two components whose values are easily seen to be

$$h_\theta = V_\theta / E_0. \tag{18}$$

$$h_\phi = V_\phi / E_1. \tag{19}$$

Thus the transmitted field, when the antenna is used for transmission is[8]

$$E = j \frac{Z_0 I h}{2 \lambda r} \epsilon^{-jkr}. \tag{20}$$

A careful distinction must be made between the vector effective length $\boldsymbol{h}$, and the effective length $l$ used in (6) through (17). The effective length $l$ is the value of the effective length for the antenna when the conducting cylinder is absent, whereas the vector effective length $\boldsymbol{h}$ is its value when the conducting cylinder is present.

When the actual expressions for the potentials $\Pi_z$ and $\Pi_z^*$ are inserted in (6) through (17), and use is made of (18) and (19), it is found that the components of the transmitted field can be written in the forms

$$h_\theta = l K_1 \sum_{n=0}^{\infty} j^n \{ A_n Se_n(s, \cos \phi_0) + B_n So_n(s, \cos \phi_0) \} \tag{21}$$

$$h_\phi = l K_2 \sum_{n=0}^{\infty} j^n \{ C_n Se_n(s, \cos \phi_0) + D_n So_n(s, \cos \phi_0) \}. \tag{22}$$

The coefficients $A_n$, $B_n$, $C_n$, $D_n$, $K_1$, and $K_2$ are listed in Table I. $Se_n$ and $So_n$ are the even and odd angular Mathieu functions and the angle $\phi_0$ is shown in Fig. 2; and $s = k^2 d^2 \sin^2 \theta_0$.

The procedure in calculating the pattern of a given antenna near an elliptic cylinder is as follows: First, the distance $2d$ between the foci of the elliptical surface is determined and the parameter $s$ calculated for the desired direction $\theta_0$. The separation constants of Mathieu's equation[9] $be_n$ and $bo_n$, and the Fourier coefficients $De_m$ and $Do_m$ are obtained from tables[10,11] or calculated.[12] The radial Mathieu functions listed in Table I are then calculated for the value $u = u_0$ defining the cylindrical surface, and for $u = u_1$ defining the location of the antenna. The values of the coefficients $K_1$, $K_2$, $A_n$, $B_n$, $C_n$, and $D_n$ are then calculated as listed in Table I. The patterns, as functions of the azimuthal angle $\phi_0$ can then be calculated using the series (21) and (22).

## ANTENNAS MOUNTED ON A CIRCULAR CYLINDER

A circular cylinder can be considered to be a limiting form of an elliptical cylinder, so the patterns of dipole

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[7] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., p. 32; 1941.

[8] G. Sinclair, "The transmission and reception of elliptically polarized waves," Proc. I.R.E., vol. 38, pp. 148–151; February, 1950. Correction, Proc. I.R.E., vol. 38, p. 1216; October, 1950.

[9] The notation for the Fourier coefficients and the separation constants was suggested by G. Blanch of the National Bureau of Standards, and conforms to that used in a forthcoming set of tables of these quantities. In the notation of footnote reference 10, $De_m = D_m{}^n$, $Do_m = F_m{}^n$, $be_n = b_n$, $bo_n = b_n'$, $s = c^2$.

[10] J. A. Stratton, P. M. Morse, L. J. Chu, and R. A. Hutner, "Elliptic Cylinder and Spheroidal Wave Functions," John Wiley and Sons, Inc., New York, N. Y.; 1941.

[11] "Table of Characteristic Values of Mathieu's Differential Equation; NDRC Applied Mathematics Panel Report No. 165-1R;" September, 1945.

[12] G. Blanch, "On the computation of mathieu functions," *Jour. Math. Phys.*, vol. 25, February 1946.

TABLE I

ANTENNAS ON AN ELLIPTIC CYLINDER

| Antenna | $K_1$ | $K_2$ | $A_n$ | $B_n$ | $C_n$ | $D_n$ |
|---|---|---|---|---|---|---|
| *Radial dipole* | $\frac{-j\sqrt{8\pi}\cos\theta_0}{kh_1\sin\theta_0}$ | $\frac{j\sqrt{8\pi}}{kh_1\sin\theta_0}$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_nHe_n^{(2)\prime}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_nHo_n^{(2)\prime}(s,\xi_1)]$ | $\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$ | $\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$ |
| *Circumferential dipole* | $\frac{-j\sqrt{8\pi}\cos\theta_0}{kh_1\sin\theta_0}$ | $\frac{-j\sqrt{8\pi}}{kh_1\sin\theta_0}$ | $\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$ | $\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_n'He_n^{(2)\prime}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_n'Ho_n^{(2)\prime}(s,\xi_1)]$ |
| *Axial dipole* | $-\sqrt{8\pi}\sin\theta_0$ | $0$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$ | $0$ | $0$ |
| *Loop with axis radial* | $\frac{-j\sqrt{8\pi}}{Z_0kh_1\sin\theta_0}$ | $\frac{-j\sqrt{8\pi}\cos\theta_0}{Z_0kh_1\sin\theta_0}$ | $\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_nHe_n^{(2)}(s,\xi_1)]$ | $\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_nHo_n^{(2)}(s,\xi_1)]$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_n'He_n^{(2)\prime}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_n'Ho_n^{(2)\prime}(s,\xi_1)]$ |
| *Loop with axis circumferential* | $\frac{j\sqrt{8\pi}}{Z_0kh_1\sin\theta_0}$ | $\frac{-j\sqrt{8\pi}\cos\theta_0}{Z_0kh_1\sin\theta_0}$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n'(s,\xi_1)+c_nHe_n^{(2)\prime}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n'(s,\xi_1)+d_nHo_n^{(2)\prime}(s,\xi_1)]$ | $\frac{Se_n'(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$ | $\frac{So_n'(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$ |
| *Loop with axis axial* | $0$ | $\frac{-\sqrt{8\pi}\sin\theta_0}{Z_0}$ | $0$ | $0$ | $\frac{Se_n(s,\cos v_1)}{N_n}[Je_n(s,\xi_1)+c_n'He_n^{(2)}(s,\xi_1)]$ | $\frac{So_n(s,\cos v_1)}{N_n'}[Jo_n(s,\xi_1)+d_n'Ho_n^{(2)}(s,\xi_1)]$ |

$\xi_1=\cosh u_1$ $\quad c_n=-\frac{Je_n(s,\cosh u_0)}{He_n^{(2)}(s,\cosh u_0)}$ $\quad d_n=-\frac{Jo_n(s,\cosh u_0)}{Ho_n^{(2)}(s,\cosh u_0)}$ $\quad s=k^2d^2\sin^2\theta_0$

$h_1=d\sqrt{\cosh^{2\prime}u_1-\cos^2 v_1}$ $\quad c_n'=-\frac{Je_n'(s,\cosh u_0)}{He_n^{(2)\prime}(s,\cosh u_0)}$ $\quad d_n'=-\frac{Jo_n'(s,\cosh u_0)}{Ho_n^{(2)\prime}(s,\cosh u_0)}$

and loop antennas mounted near a circular cylinder can be obtained from the above results by taking the appropriate limits. The transition to co-ordinates of a circular cylinder is given by the limits

$$d\to 0 \tag{23}$$

$$u\to\infty \tag{24}$$

in such a manner that the product $de^u$ remains finite and equal to the radius $\rho$ of the circular cylinder. The angular variable $v$ becomes the azimuthal angle $\phi$. It is, however, more convenient to obtain the expressions directly as shown by Carter.[1,2] The resulting expressions for the components of the vector effective length are Fourier series of the forms

$$h_\theta = lK_1 \sum_{n=0}^{\infty} \epsilon_n j^n \{A_n \cos n\phi_0 + nB_n \sin n\phi_0\} \qquad (25)$$

$$h_\phi = lK_2 \sum_{n=0}^{\infty} \epsilon_n j^n \{C_n \cos n\phi_0 + nD_n \sin n\phi_0\}, \qquad (26)$$

where $\epsilon_n$ is Neumann's number ($\epsilon_0 = 1$, otherwise $\epsilon_n = 2$), and the coefficients $K_1$, $K_2$, $A_n$, $B_n$, $C_n$ and $D_n$ are listed in Table II. The results given in Table II are due to Carter[1,2] and are repeated here for convenience and completeness. The antenna is located at the point $P_1(b, \phi_1, z_1)$, and there is no loss in generality in assuming that $\phi_1 = z_1 = 0$. The radius of the perfectly conducting infinite circular cylinder is $\rho_0 = a$ (see Fig. 3).

## EXPERIMENTAL VERIFICATION OF THE CALCULATED PATTERNS

A number of measurements have been made to test the accuracy of the calculated patterns with the results shown in Figs. 4 and 5. In all cases the antenna was mounted directly on the surface of the cylinder so that $u_1 = u_0$. Fig. 4 shows the patterns for a very short radial stub antenna for two different locations, while Fig. 5 shows the patterns for a loop antenna. The poorer agreement between measurements and calculations in the case of the loop antenna is attributed to the finite dimensions of the loop used in the measurements. Measurements verifying the accuracy of the calculations for antennas mounted near a circular cylinder have been described previously.[1,2,13]

## ANTENNAS OF ARBITRARY CONFIGURATIONS

The results derived above for the patterns of antennas located near circular and elliptic cylinders are actually the expressions for the distant fields radiated by

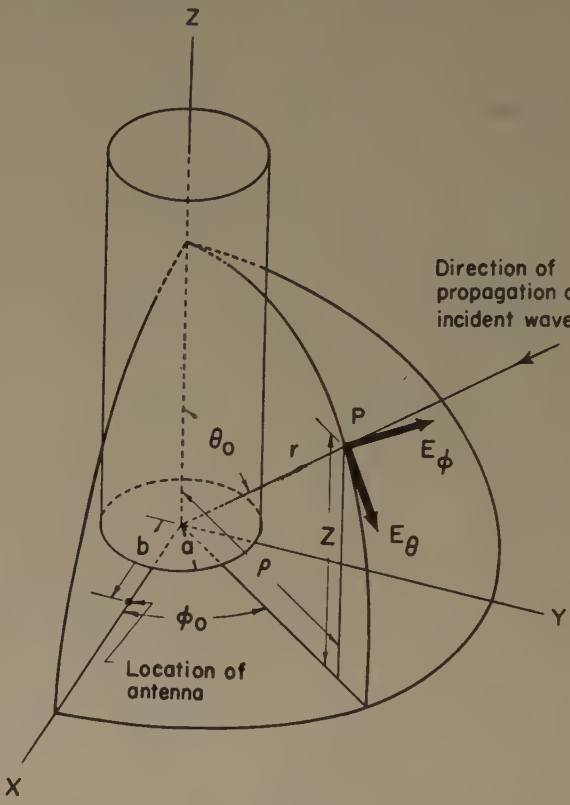

Fig. 3—Co-ordinate system for the circular cylinder.

TABLE II

ANTENNAS ON A CIRCULAR CYLINDER

| Antennas | $K_1$ | $K_2$ | $A_n$ | $B_n$ | $C_n$ | $D_n$ |
|---|---|---|---|---|---|---|
| *Radial dipole* | $-j \cos \theta_0$ | $-j \dfrac{1}{w}$ | $J_n'(w) + c_n H_n^{(2)\prime}(w)$ | $0$ | $0$ | $J_n(w) + c_n' H_n^{(2)}(w)$ |
| *Circumferential dipole* | $j \dfrac{\cos \theta_0}{w}$ | $-j$ | $0$ | $J_n(w) + c_n H_n^{(2)}(w)$ | $J_n'(w) + c_n' H_n^{(2)\prime}(w)$ | $0$ |
| *Axial dipole* | $-\sin \theta_0$ | $0$ | $J_n(w) + c_n H_n^{(2)}(w)$ | $0$ | $0$ | $0$ |
| *Loop with axis radial* | $-\dfrac{1}{jZ_0 w}$ | $-j \dfrac{\cos \theta_0}{Z_0}$ | $0$ | $J_n(w) + c_n H_n^{(2)}(w)$ | $J_n'(w) + c_n' H^{(2)\prime}(w)$ | $0$ |
| *Loop with axis circumferential* | $-\dfrac{1}{jZ_0}$ | $j \dfrac{\cos \theta_0}{Z_0 w}$ | $J_n'(w) + c_n H_n^{(2)\prime}(w)$ | $0$ | $0$ | $J_n(w) + c_n' H_n^{(2)}(w)$ |
| *Loop with axis axial* | $0$ | $-\dfrac{\sin \theta_0}{Z_0}$ | $0$ | $0$ | $J_n(w) + c_n' H_n^{(2)}(w)$ | $0$ |

$$w = kb \sin \theta_0 \qquad c_n = -\frac{J_n(ka \sin \theta_0)}{H_n^{(2)}(ka \sin \theta_0)} \qquad c_n' = -\frac{J_n'(ka \sin \theta_0)}{H_n^{(2)\prime}(ka \sin \theta_0)}$$

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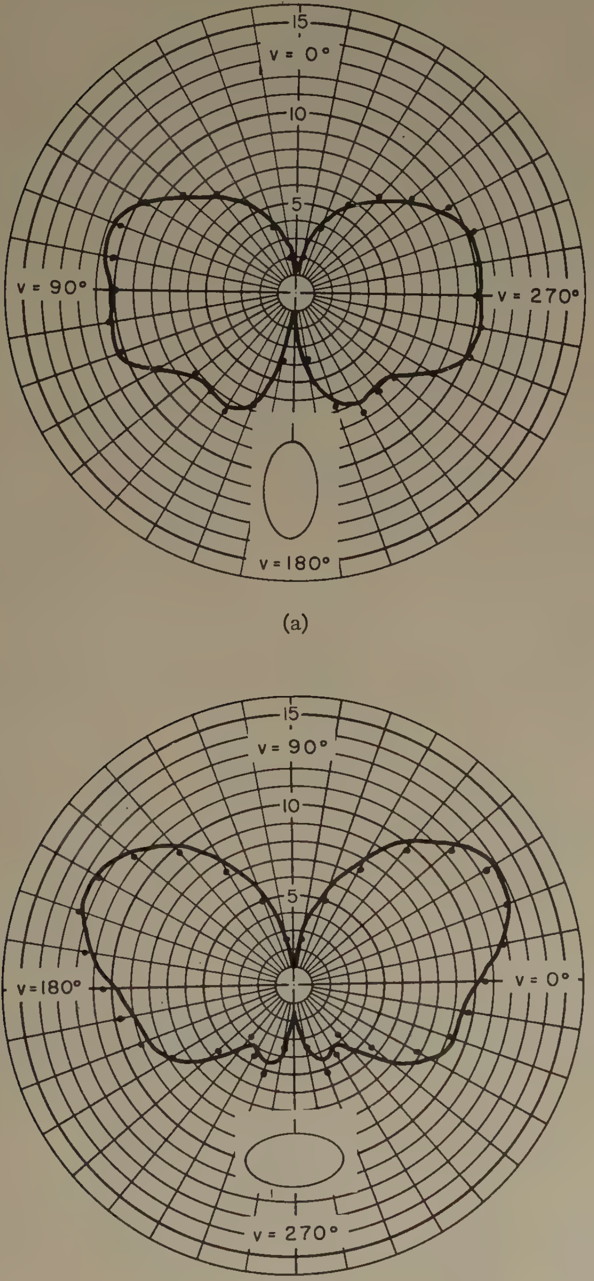
[13] G. Sinclair, E. C. Jordan, and E. W. Vaughan, "Measurement of aircraft-antenna patterns using models," PROC. I.R.E., vol. 35, pp. 1451–1462; December, 1947.

infinitesimal elements of electric and magnetic currents. Therefore, it is possible to use these results to determine the patterns of more complicated antennas when the distributions of current or voltage along the antenna are known since it is merely necessary to sum the contributions to the total field from each part of the antenna. It is possible to carry out the integrations involved in this calculation in certain cases, as, for example, when the current or voltage can be assumed distributed sinusoidally.



(a)

(b)

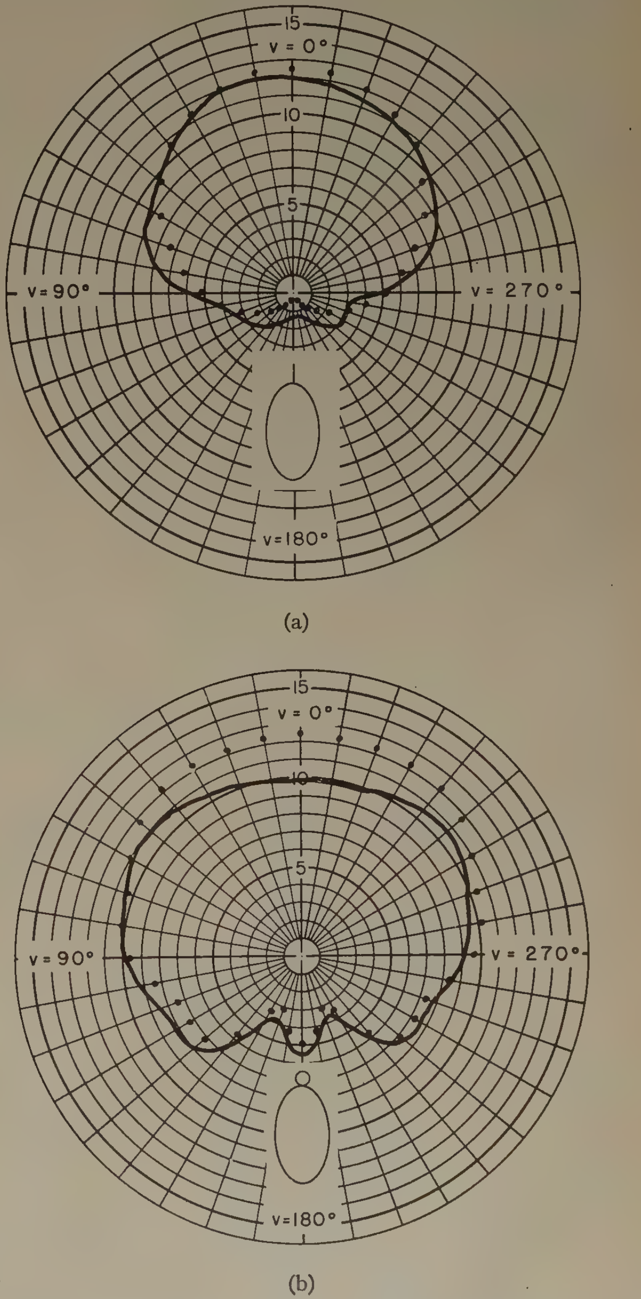
Fig. 4—Comparison of measured and calculated patterns for a short radial stub antenna mounted on an elliptic cylinder of semimajor axis $a'=0.637\lambda$, and semiminor axis $b'=0.390\lambda$. Curves measured at 1,500 mc and points calculated. (a) Stub located at $v_1=0$, $u_1=u_0$. (b) Stub located at $v_1=\pi/2$, $u_1=u_0$.

## CONCLUSIONS

Expressions have been obtained for the distant field radiated by small dipole and loop antennas located on or near cylinders of elliptical cross section. Arrays of such antennas can be treated by the superposition of the fields due to the individual radiators.[1]

## ACKNOWLEDGEMENTS

The measurements shown in Figs. 4 and 5 were made by R. A. Fouty; the calculations, by Miss Frances M. Nichols. Some assistance in calculating the Mathieu functions was received from G. Blanch of the National Bureau of Standards.



(a)

(b)

Fig. 5—Similar to Fig. 4, but for a loop antenna $u_1=u_0$, $v_1=0$. (a) Loop axis directed circumferentially. (b) Loop axis directed parallel to axis of cylinder.

## APPENDIX

### *Diffraction of a Plane Wave Around an Elliptic Cylinder*

The diffraction of a plane wave around an elliptic cylinder is readily solved by expressing the vector wave equation in elliptic cylinder co-ordinates.[14] The equation for the $z$ component of the Hertzian potential $\mathbf{\Pi}$ is

$$\frac{1}{h_1^2}\left\{\frac{\partial^2\Pi_z}{\partial u^2}+\frac{\partial^2\Pi_z}{\partial v^2}\right\}+\frac{\partial^2\Pi_z}{\partial z^2}+k^2\Pi_z=0, \quad (27)$$

where the metrical coefficient is

$$h_1^2 = d^2(\cosh^2 u - \cos^2 v). \quad (28)$$

This equation is readily solved by separation, that is, by assuming a solution of the form

$$\Pi_z = f_1(u)f_2(v)f_3(z), \quad (29)$$

where the functions $f_i$ are functions only of the single variable shown for each.

The function $f_3(z)$ is the exponential function

$$f_3(z) = e^{\pm j\gamma z}, \quad (30)$$

where $\gamma$ is a constant resulting from the separation of the differential equation.

For the function $f_2(v)$, the periodic even and odd angular Mathieu functions are chosen

$$Se_n(s, \cos v) = \sum_{m=0,1}^{\infty}{}' De_m \cos mv \quad (31)$$

$$So_n(s, \cos v) = \sum_{m=1,2}^{\infty}{}' Do_m \sin mv, \quad (32)$$

where

$$s = d^2(k^2 - \gamma^2) \quad (33)$$

and the prime on the summation sign indicates a summation over even values of $m$ if $n$ is even, and over odd values of $m$ if $n$ is odd.

The radial Mathieu functions $f_1(u)$ associated with the even and odd angular Mathieu functions are[15]

$$Ze_n(s, \cosh u) = \sqrt{\frac{\pi}{2}} \sum_{m=0,1}^{\infty}{}' j^{n-m} De_m Z_m(s \cosh u) \quad (34)$$

$$Zo_n(s, \cosh u) = \sqrt{\frac{\pi}{2}} \sum_{m=1,2}^{\infty}{}' j^{n-m} m Do_m Z_m(s \cosh u), \quad (35)$$

where $Z_m(s)$ is any cylinder function.

Since any linear combination of solutions is still a solution of (27), a complete solution is therefore

$$\Pi_z = \sum_{n=0}^{\infty} \{c_n Se_n(s, \cos v) Ze_n(s, \cosh u)$$
$$+ d_n So_n(s, \cos v) Zo_n(s, \cosh u)\} \times e^{\pm\gamma z}. \quad (36)$$

The constant $\gamma$ and the coefficients $c_n$ and $d_n$ are to be determined by the physical conditions of the boundary value problem.

Consider a plane wave polarized parallel to the theta component of the radiated field of the antenna shown in Fig. 2. The incident plane wave is expressed by the equation

$$E^i = E_0 e^{jkr+j\omega t}, \quad (37)$$

where $r$ is distance measured from the origin of the co-ordinate system along the direction of propagation (see Fig. 2). In elliptical cylinder co-ordinates,

$$r = d \sin\theta_0 (\cosh u \cos v \cos\phi_0$$
$$+ \sinh u \sin v \sin\phi_0) + z \cos\theta_0, \quad (38)$$

where the angles $\phi_0$ and $\theta_0$ define the direction of incidence. It is readily shown that this plane wave can be derived from the following Hertzian potential having only a $z$ component:

$$\Pi_z{}^i = \frac{-E_0 e^{jkr}}{k^2 \sin\theta_0}$$
$$= \frac{-\sqrt{8\pi}\,E_0}{k^2\sin\theta_0}\sum_{n=0}^{\infty} j^n \left\{\frac{1}{N_n} Je_n(s, \cosh u)\right.$$
$$\cdot Se_n(s, \cos v) Se_n(s, \cos\phi_0)$$
$$+ \frac{1}{N_n{}'} Jo_n(s, \cosh u) So_n(s, \cos v)$$
$$\left. \cdot So_n(s, \cos\phi_0)\right\} e^{jkz\cos\theta_0}, \quad (39)$$

where

$$N_n = \pi\left\{2(De_0)^2 + \sum_{m=2}^{\infty}{}'(De_m)^2\right\} \text{ when } n \text{ is even} \quad (40)$$

$$N_n = \pi \sum_{m=1}^{\infty}{}'(De_m)^2 \text{ when } n \text{ is odd} \quad (41)$$

$$N_n{}' = \pi \sum_{m=1,2}^{\infty}{}' (Do_m)^2 \quad (42)$$

$$s = k^2 d^2 \sin^2\theta_0 \quad (43)$$

$$\gamma = k \cos\theta_0. \quad (44)$$

It is assumed that the boundary conditions can be satisfied by choosing for the reflected field the following potential:

$$\Pi_z{}^r = \frac{-\sqrt{8\pi}\,E_0}{k^2\sin\theta_0}\sum_{n=0}^{\infty} j^n \left\{\frac{c_n}{N_n} He_n{}^{(2)}(s, \cosh u)\right.$$
$$\cdot Se_n(s, \cos v) Se_n(s, \cos\phi_0)$$
$$+ \frac{d_n}{N_n{}'} Ho_n{}^{(2)}(s, \cosh u) So_n(s, \cos v)$$
$$\left. \cdot So_n(s, \cos\phi_0)\right\} e^{jkz\cos\theta_0}, \quad (45)$$

where $He_n{}^{(2)}(s, \cosh u)$ and $Ho_n{}^{(2)}(s, \cosh u)$ are radial

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[14] See pp. 375–386 of footnote reference 7.

[15] The notation employed for the general Mathieu function is an obvious extension of the notation used in footnote reference 10.

Mathieu functions of the fourth kind, and are the appropriate functions for ensuring that the reflected field consists only of outward travelling waves.[16]

The total field can be derived from the potential

$$\Pi_z = \Pi_z^i + \Pi_z^r. \tag{46}$$

The components of the total field are found from the expressions

$$\mathbf{E} = \text{curl curl } \mathbf{\Pi} \tag{47}$$

$$\mathbf{H} = j\omega\epsilon \text{ curl } \mathbf{\Pi}. \tag{48}$$

Thus the tangential components of the electric vector are

$$E_v = \frac{1}{h_1}\frac{\partial^2 \Pi_z}{\partial v \partial z} \tag{49}$$

$$E_z = k^2 \sin^2 \theta_0 \Pi_z. \tag{50}$$

Since the boundary conditions require that the tangential components of the electric vector must vanish on the conducting surface $u = u_0$, it is apparent that $c_n$ and $d_n$ must have the values

$$c_n = -\frac{Je_n(s, \cosh u_0)}{He_n^{(2)}(s, \cosh u_0)} \tag{51}$$

$$d_n = -\frac{Jo_n(s, \cosh u_0)}{Ho_n^{(2)}(s, \cosh u_0)}. \tag{52}$$

Thus the components of the total field can be found.

For an incident plane wave polarized parallel to the phi component of the pattern, the expression for the plane wave is

$$E^i = E_1 e^{jkr + j\omega t}. \tag{53}$$

This field can be derived from the potential

$$\pi_z^{*i} = \frac{-E_1 e^{jkr}}{Z_0 k^2 \sin \theta_0}. \tag{54}$$

Expressing this potential in terms of the wave functions and assuming a suitable form for the potential of the reflected field, it is found that the total field can be derived from the potential

$$\begin{aligned} \Pi_z^* &= \Pi_z^{*i} + \Pi_z^{*r} \\ &= \frac{-\sqrt{8\pi}\, E_1}{Z_0 k^2 \sin \theta_0} \sum_{n=0}^{\infty} j^n \Bigg\{ \frac{1}{N_n} [Je_n(s, \cosh u) \\ &\quad + c_n' He_n^{(2)}(s, \cosh u)] Se_n(s, \cos v) Se_n(s, \cos \phi_0) \\ &\quad + \frac{1}{N_n'} [Jo_n(s, \cosh u) + d_n' Ho_n^{(2)}(s, \cosh u)] \\ &\qquad \cdot So_n(s, \cos v) So_n(s, \cos \phi_0) \Bigg\} e^{jkz \cos \theta_0}, \end{aligned} \tag{55}$$

where $c_n'$ and $d_n'$ are to be determined by the boundary conditions, namely, the vanishing of the tangential components of the electric vector. The components of the field due to this potential are found from

$$\mathbf{E} = -j\omega\mu \text{ curl } \mathbf{\Pi}^* \tag{56}$$

$$\mathbf{H} = \text{curl curl } \mathbf{\Pi}^*. \tag{57}$$

The tangential components of the electric vector are

$$E_v = j\frac{\omega\mu}{h_1}\frac{\partial \Pi_z^*}{\partial u} \tag{58}$$

$$E_z = 0. \tag{59}$$

The boundary conditions are satisfied by choosing

$$c_n' = -\frac{Je_n'(s, \cosh u_0)}{He_n^{(2)\prime}(s, \cosh u_0)} \tag{60}$$

$$d_n' = -\frac{Jo_n'(s, \cosh u_0)}{Ho_n^{(2)\prime}(s, \cosh u_0)} \tag{61}$$

where the primed Mathieu functions mean the derivatives with respect to the argument $u$, that is, for example,

$$Je_n'(s, \cosh u_0) = \left[\frac{d}{du} Je_n(s, \cosh u)\right]_{u=u_0}. \tag{62}$$

In Table I, the prime on the angular Mathieu functions means, for example,

$$Se_n'(s, \cos v_1) = \left[\frac{d}{dv} Se_n(s, \cos v)\right]_{v=v_1}. \tag{63}$$

### List of Symbols

$A_n$ = series coefficient (see Tables I and II)
$B_n$ = series coefficient (see Tables I and II)
$C_n$ = series coefficient (see Tables I and II)
$D_n$ = series coefficient (see Tables I and II)
$De_m$ = Fourier coefficient for even Mathieu functions
$Do_m$ = Fourier coefficient for odd Mathieu functions
$\mathbf{E}$ = electric vector
$\mathbf{H}$ = magnetic vector
$He_n^{(2)}(s, \cosh u)$ = even radial Mathieu function of fourth kind
$Ho_n^{(2)}(s, \cosh u)$ = odd radial Mathieu function of fourth kind
$Je_n(s, \cosh u)$ = even radial Mathieu function of first kind
$Jo_n(s, \cosh u)$ = odd radial Mathieu function of first kind
$K_1$ = see Tables I and II
$K_2$ = see Tables I and II
$N_n$ = normalization factor for $Se_n(s, \cos v)$
$N_n'$ = normalization factor for $So_n(s, \cos v)$
$S$ = area of loop antenna in square meters
$Se_n(s, \cos v)$ = even angular Mathieu function
$So_n(s, \cos v)$ = odd angular Mathieu function
$V$ = open-circuit terminal voltage
$Z_0$ = intrinsic impedance of free space (120π ohms)

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[16] Note that $e^{+j\omega t}$ has been assumed for the time dependence of the field.

$a$ = radius of perfectly conducting circular cylinder
$a'$ = semimajor axis of an ellipse
$b$ = radial distance to location of antenna in circular cylindrical co-ordinates
$b'$ = semiminor axis of an ellipse
$c_n$ = series coefficient (see (51))
$c_n'$ = series coefficient (see (60))
$d$ = half the distance between foci of the ellipses
$d_n$ = series coefficient (see (52))
$d_n'$ = series coefficient (see (61))
$h_1$ = metrical coefficient $h_1 = d\sqrt{\cosh^2 u - \cos^2 v}$
$k$ = propagation constant, $k = 2\pi/\lambda$
$l$ = effective length of dipole or loop antenna
$r$ = radial distance from origin
$s = k^2 d^2 \sin^2\theta_0$
$u$ = radial co-ordinate in elliptic cylinder co-ordinates
$u_0$ = radial co-ordinate for surface of perfectly conducting elliptical cylinder
$u_1$ = radial co-ordinate of location of antenna
$v$ = angular co-ordinate in elliptic cylinder co-ordinates
$v_1$ = angular co-ordinate of location of antenna
$w = kb \sin\theta_0$ (see Table II)
$z$ = axial co-ordinate
$\Pi$ = Hertzian vector potential
$\Pi^*$ = Hertzian vector potential
$\gamma$ = separation constant, $\gamma = k \cos\theta_0$
$\epsilon$ = dielectric constant of free space, $\epsilon = 10^{-9}/36\pi$
$\epsilon_n$ = Neumann's number ($\epsilon_0 = 1$; $\epsilon_n = 2$ if $n \neq 0$)
$\theta_0$ = angle between $z$ axis and direction of propagation
$\lambda$ = wavelength in free space
$\mu$ = permeability of free space, $\mu = 4\pi \times 10^{-7}$
$\xi = \xi_1 = \cosh u_1$ (see Table I)
$\rho$ = radial co-ordinate in circular cylindrical co-ordinates
$\phi$ = azimuthal angular co-ordinate
$\phi_0$ = azimuthal angle of direction of propagation.

# Current Distributions on Helical Antennas*

JAMES A. MARSH†, ASSOCIATE, IRE

*Summary*—**The current distribution on a uniform circular helix has been measured over a frequency band extending from 600 to 1,700 mc. Although the measured distributions are highly complex functions of distance and also change in an anomalous manner with frequency, it is possible to analyze the distributions in terms of traveling waves associated with three different transmission modes on the helix. The relative amplitude functions of these traveling waves, as well as their associated phase velocities, are approximated. Current distributions which have been calculated by superposing three or more traveling waves are in good agreement with the measured data.**

## I. INTRODUCTION

IN RECENT years considerable effort has been spent in analyzing the propagation of electrical energy along a helical conductor. Many applications of the helix have been found. One of these, discovered by Kraus,[1–5] employs the helix as an end-fire antenna. In this application the pitch angle may lie between 10 and 20 degrees while the circumference of the imaginary cylinder having diameter $D$ (see Fig. 1) is approximately 0.80 to 1.3 wavelengths. In this range of frequencies a helix, of at least a few turns, radiates approximately circularly polarized waves in the direction of its axis. The radiation pattern of this antenna has a beamwidth which may be made quite narrow by employing a sufficiently large number of turns. The helix, when used in this manner, may be called a "helical beam antenna" and is said to be radiating in the "axial mode."

Although the current distributions of many types of antennas are well known, only a few distributions have been reported for the helical antenna.[2] The current

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[1] J. D. Kraus, "Helical beam antenna," *Electronics*, vol. 20, pp. 109–111; April, 1947.

[2] J. D. Kraus and J. C. Williamson, "Characteristics of helical antennas radiating in the axial mode," *Jour. Appl. Phys.*, vol. 19, pp. 87–96; January; 1948.

[3] O. J. Glasser and J. D. Kraus, "Measured impedances of helical beam antennas," *Jour. Appl. Phys.*, vol. 19, pp. 193–197; February; 1948.

[4] J. D. Kraus, "Helical beam antennas for wide-band applications," PROC. I.R.E., vol. 36, pp. 1236–1242; October, 1948.

[5] J. D. Kraus, "The helical antenna," PROC. I.R.E., vol. 37, pp. 263–272; March, 1949.

distribution on the helix is complex, and varies in an anomalous manner with changes in frequency. Accordingly, a systematic set of measurements was undertaken in order to observe the current distribution both as a function of distance along the helix and also as a function of the frequency. Examples of these measured distributions and their analysis in terms of traveling waves on the helix are presented in this paper.

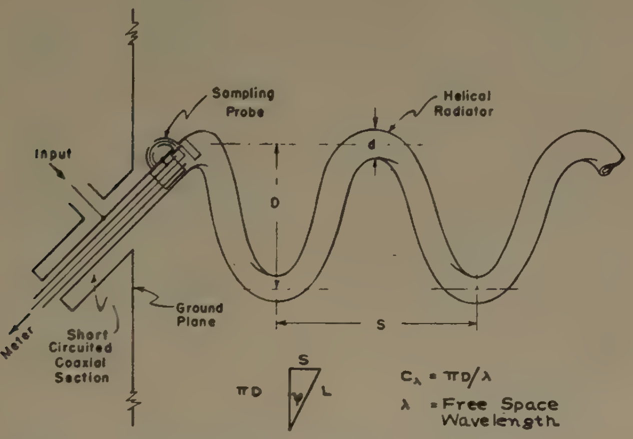
Sampling Probe
Helical Radiator
Input
d
D
Ground Plane
Meter
Short Circuited Coaxial Section
S
S
πD
L
ψ
$C_\lambda = \pi D/\lambda$
λ = Free Space Wavelength

Fig. 1—Sketch of feed and sampling system used in the investigation of the helix.

All of the measurements were made in the uhf range on helices made from copper cylindrical tubing having small outer diameters in terms of a wavelength. Therefore, the assumption is made that the current on the helix is a one-dimensional surface current flowing only in the direction of the conductor axis. This assumption permits the use of the method of feeding and sampling shown in Figs. 1 and 2.[6] The hollow-center conductor of a short-circuited coaxial section is extended and wound in the shape of a uniform circular helix. The helix is excited by feeding energy through the side of the coaxial section. The current on the helix is sampled by means of a balanced loop which is positioned in an axial slot in the radiator. The energy sampled by the loop is fed back through the inside of the helix on a coaxial line to the measuring equipment located behind the ground plane.

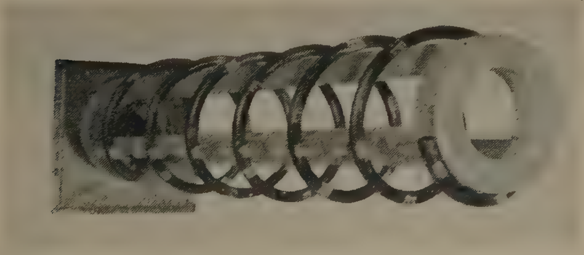

Fig. 2—Six-turn uniform circular helix, pitch angle $\psi = 12.6°$, mean diameter $D = 8.61$ cm.

Since the currents on the helix vary harmonically with time, must satisfy Maxwell's equations, and are assumed to be one-dimensional, it is possible to obtain expressions for the currents by the superposition of one-dimensional traveling waves.

The analysis of a measured current distribution in terms of traveling waves consists of assigning a relative amplitude function and a relative phase velocity to each of the component waves. The analysis becomes difficult for the present case, since there are at least three traveling waves of significant magnitude which, in general, have different phase velocities. For simplicity, the amplitude of the component waves are assumed to be either constant or exponentially attenuated. Therefore, each component wave can be expressed in the form:

$$I = I_0 e^{-\alpha x + j(\omega t \mp \beta x)},$$

where $\alpha$ is the attenuation constant, $\omega$ is the angular frequency, and $\beta$ is the phase constant. The phase velocity of such a wave is usually defined[7] as $v = \omega/\beta$ and is independent of the attenuation constant $\alpha$. In practical systems, however, the existence of a single traveling wave of the form above is rarely encountered. The current distributions that are usually measured consist of two or more traveling waves, and only the sum total of these currents is measurable.

It is common practice to measure the relative phase and amplitude of the total current as functions of distance and plot these data, choosing a convenient reference point in the system for the co-ordinate, in a manner as shown in Figs. 3 to 5. The relative amplitude curves are labeled $|I(x)|^2$ and the relative phase curves are labeled $\phi$. The angle $\phi$, essentially a lag angle for waves traveling to the right, is plotted positively for convenience. Since the angular frequency $\omega$ is readily determined in any practical measurement, it is only necessary to obtain the wave number $\beta$ to find the phase velocity. If there were but one wave traveling with a constant velocity in the system under consideration, the $\phi(x)$ curve would be linear with a slope equal to $\beta$. The determination of the phase velocity for this case is straightforward. However, where more than one wave is present, the curve is not linear and the determination of $\beta$ must be made indirectly. That is, an average slope of the $\phi$ curve, or some other method which uses the distance between successive points which are 360° apart must be employed to obtain the phase velocity.

A generalized phase velocity can be defined for the case where more than one traveling wave is present in the system by letting the slope $d\phi/dx = \beta(x)$ and defining $\hat{v} = -\omega/\beta(x)$, where $\hat{v}$ is called the generalized

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[6] This method has been used by a number of investigators; see for example, G. Barzilai, "Experimental determination of the distribution of current and charge along cylindrical antennas," Proc. I.R.E., vol. 37, pp. 825–829; July, 1949. Also, M. Aronoff, "Measured phase velocity and current distribution characteristics of helical antennas radiating in the beam mode," an unpublished thesis for the degree of M.Sc., Ohio State University, 1948.

[7] J. A. Stratton, "Electromagnetic Theory," 1st ed., McGraw-Hill Book Co., Inc., New York, N. Y.; 1941.

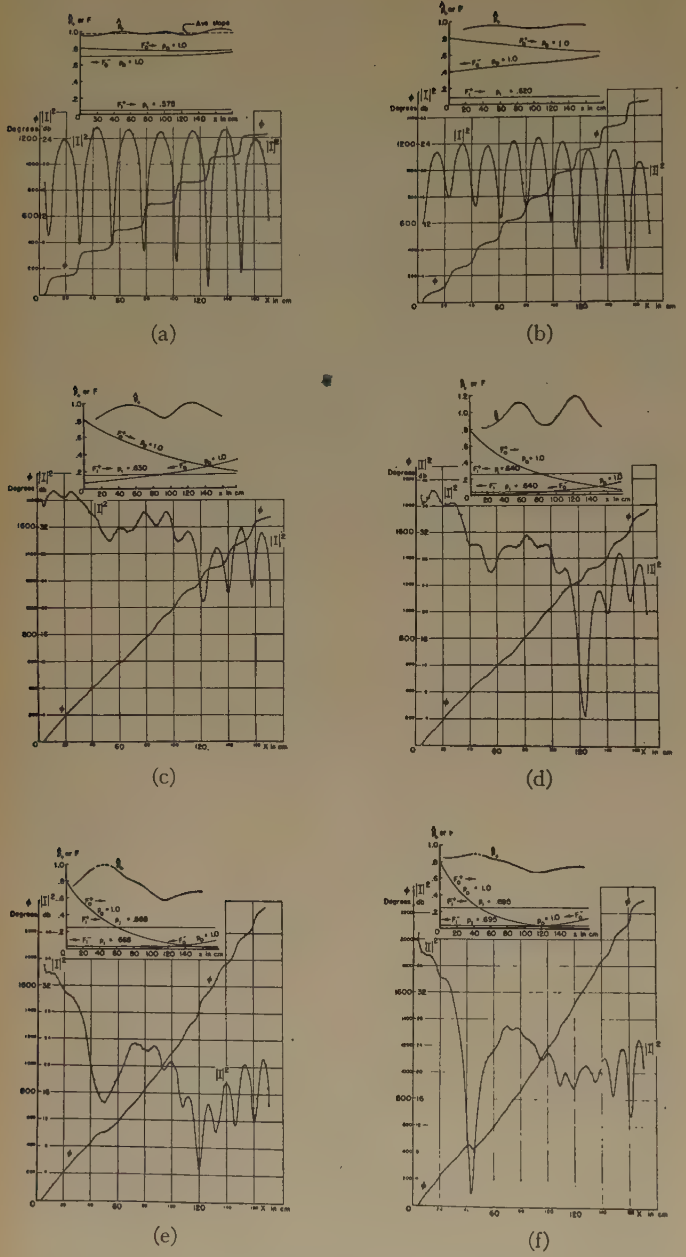
(a) (b) (c) (d) (e) (f)

Fig. 3—Measured current distributions on the helix as a function of $C_\lambda$, the circumference in wavelengths:
(a) $C_\lambda=0.542$ (b) $C_\lambda=0.665$
(c) $C_\lambda=0.691$ (d) $C_\lambda=0.715$
(e) $C_\lambda=0.774$ (f) $C_\lambda=0.855$.

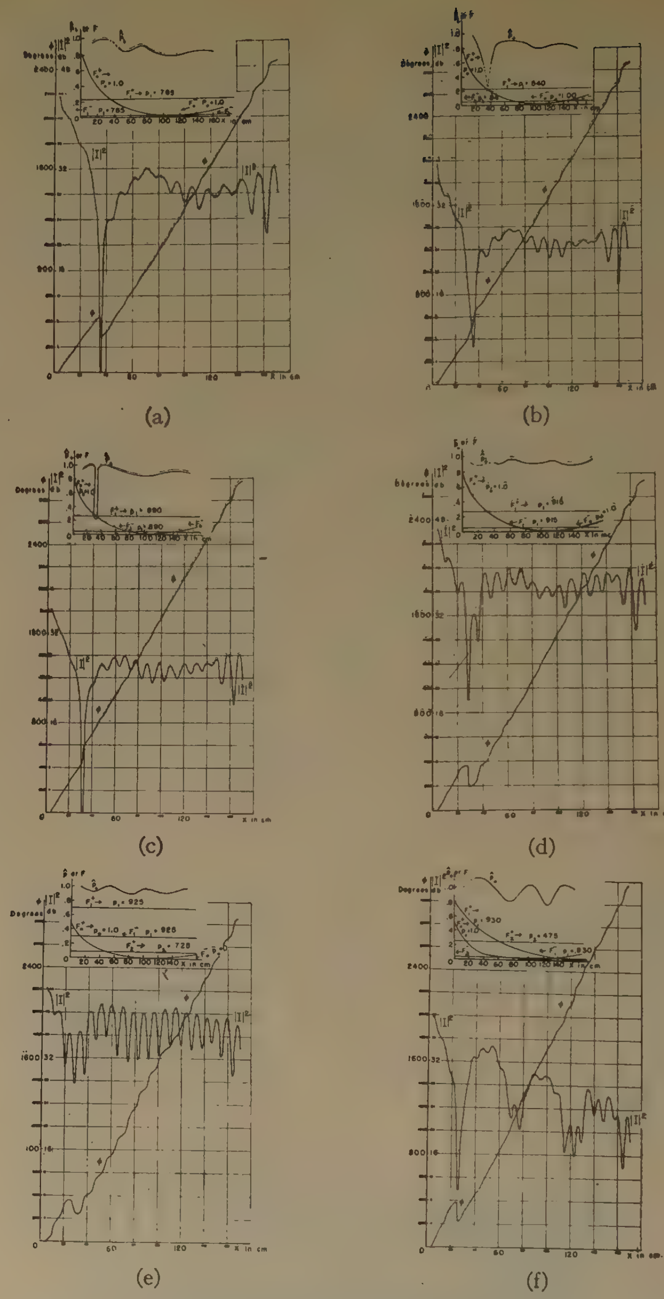
(a) (b) (c) (d) (e) (f)

Fig. 4—Measured current distributions on the helix as a function of $C_\lambda$, the circumference in wavelengths:
(a) $C_\lambda=1.01$ (b) $C_\lambda=1.09$
(c) $C_\lambda=1.175$ (d) $C_\lambda=1.26$
(e) $C_\lambda=1.36$ (f) $C_\lambda=1.50$.

phase velocity, and the negative sign is necessary because of the convention adopted in plotting $\phi$ positive for a wave traveling in the positive $x$ direction. Such a generalization leads to many interesting interpretations of phenomena often observed.[8] If $\omega$ and $\phi(x)$ are measured or known mathematically, the quantity $\hat{v}$ may be found graphically or calculated directly. Since the phase velocity relative to the velocity of light is usually used to describe the traveling waves in the system, the data are presented in the form $\hat{p}=\hat{v}/c$, where $c$ is the velocity of light and $\hat{p}$ is called the generalized relative phase velocity. Plots of $\hat{p}_+$, the generalized relative phase velocity for those waves traveling in the positive $x$ direction only, may be found in Figs. 3 and 4.

## II. MEASURED DISTRIBUTIONS

The antenna investigated (see Fig. 2) is a six-turn helix which has a pitch angle $\Psi=12.6°$, and a mean diameter $D=8.61$ cm. This antenna is mounted on a three-foot-square ground plane which is supported so that the helix is horizontal and about six wavelengths above ground. The current distribution was obtained as a function of distance measured along the slot in the

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[8] For a more complete discussion, see J. A. Marsh, "A study of phase velocity on long cylindrical conductors," presented, Graduate School, Ohio State University, 1949. Presented, 1950 National IRE Convention, New York, N. Y., March 7, 1950. Published as a special report, The Ohio State University Research Foundation, February 28, 1950.

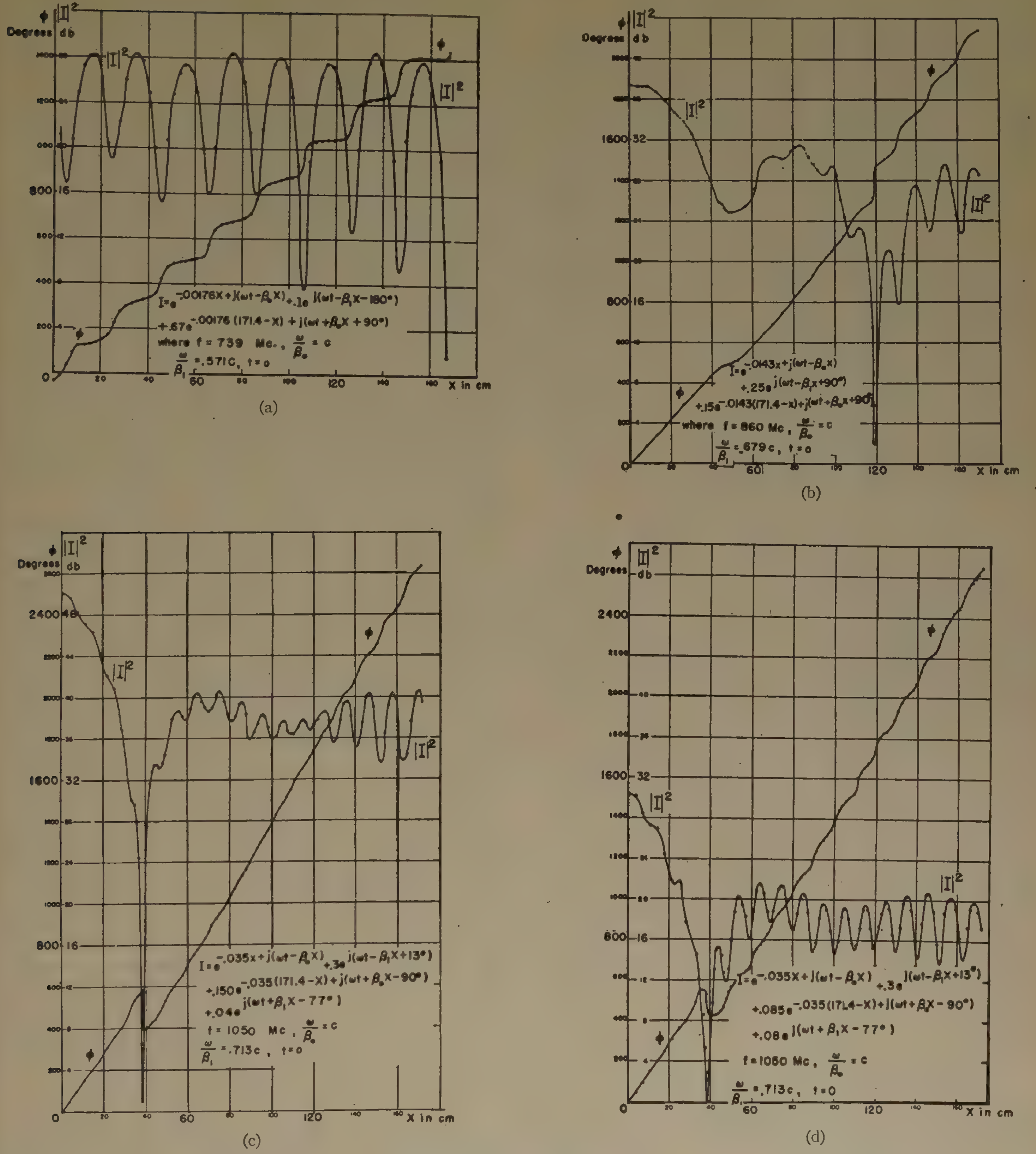

Fig. 5—Calculated current distributions on the helix:
(a) To mode region (compare with Fig. 3(b)).
(b) Transition region from $T_0$ to $T_1$ modes (compare with Fig. 3(d)).
(c) $T_1$ mode region (compare with Fig. 4(a)).
(d) Transition region from $T_1$ to $T_2$ modes (Compare with Fig. 4(d) and 4(e)).

hollow tubing using the ground plane as a reference.

The measured current distribution on the helix for the range of frequencies 602 to 1,667 mc may be seen in Figs. 3 and 4. The distributions are labeled in terms of the parameter $C_\lambda$, which is defined in Fig. 1 as the circumference in free-space wavelengths of the cylinder with diameter $D$. The data[9] are arranged in order of increasing $C_\lambda$. In each of the distributions shown in

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[9] The magnitude of the current $I$ is expressed in decibels relative to an arbitrary reference current $I_0$ by db $=10 \log (I^2/I_0^2)$.

Figs. 3 and 4 there is an insert which contains a portion of the analysis made of the data for that distribution. These inserts show an approximate plot of the generalized relative phase velocity, $\hat{p}_+$ for the waves which travel only in the positive $x$-direction away from the ground plane. The other curves, shown in the inserts, are plots of the amplitude or $F$ functions for the number of one-dimensional traveling waves which may be used to obtain an approximate expression for the current distribution shown. The plus and minus signs are used to show that the associated waves are traveling in the positive or negative $x$-direction, respectively. It should be stated that these curves are only initial assumptions needed to "fit" the associated distribution. Because of the number of variables involved, the actual task of fitting any one of these curves closely is tremendous. Thus, no attempt was made to "fit" a particular distribution exactly. Four distributions (see Fig. 5) were calculated to show how distributions in four different regions of $C_\lambda$ could be obtained.

## III. ANALYSIS

In the analysis of the data shown in Figs. 3 and 4, it is found that the current distributions can be interpreted in terms of traveling waves associated with three different modes of transmission on the helix. These modes[5] are distinguished by means of the symbols $T_0$, $T_1$, and $T_2$. The $T_0$ mode of transmission on the helix is found to predominate when $C_\lambda < 0.675$. This is the mode of transmission which is encountered in the traveling wave tube. The $T_1$ mode of transmission predominates when $C_\lambda$ lies in the region from 0.80 to 1.30. Since the $T_1$ mode of transmission on the helix makes possible the axial or beam mode of radiation, it is the center of interest in this investigation. The $T_2$ mode of propagation exists when $C_\lambda > 1.25$. The presence of energy in this mode is first detected at the upper frequency limit of the helical beam antenna. The subscripts 0, 1, and 2 are used in the inserts of Figs. 3 and 4 to connect the relative amplitude functions and phase velocities of the traveling waves with these transmission modes.

### A. The $T_0$ Mode Distribution

The current distribution shown in Fig. 3(a) is a typical example of the distributions found in the range of frequencies where $C_\lambda$ varies from 0.348 to 0.600. Traveling waves of current, which may be associated with the $T_0$ mode of transmission, predominate in this range of frequencies. The amplitudes of the $T_0$ traveling waves encounter only a slight attenuation and the waves are assumed to travel with the velocity of light. This results in an amplitude characteristic which has a large standing-wave ratio and a phase characteristic which is nearly a step function.

The presence of $T_1$ mode currents may be detected in the current distributions when $C_\lambda = 0.542$ and 0.665 (see Fig. 3(a), (b)) by observing the undulating nature of the $\hat{p}_+$ curves in the inserts. However, the relative importance of the $T_1$ mode currents in these two distributions is very slight.

A calculated $T_0$-mode-region current distribution is presented in Fig. 5(a). This distribution is a first attempt to approximate the measured current distribution shown in Fig. 3(b), when $C_\lambda = 0.665$. Although it is obvious from a comparison of the two distributions that refinements on the first attempt could be made, it is felt that the calculated distribution shows clearly that, by minor manipulations of the available parameters of the three traveling waves, one could obtain calculated and measured distributions which agree within experimental accuracy. The assumed current expression consists of two $T_0$-mode traveling waves which travel with the velocity of light in opposite directions and have a slight exponential amplitude attenuation. The third, $T_1$ mode, wave is assumed to be constant in amplitude and to travel with a relative velocity $p_1 = 0.571$.

### B. The Transition Region from the $T_0$ to the $T_1$ Mode

The transition region from the range of frequencies where the $T_0$ mode predominates to the range of frequencies where the $T_1$ mode predominates, occurs in the range $C_\lambda = 0.675$ to $C_\lambda = 0.800$. A study of the measured current distributions in Fig. 3 shows how this transition occurs. A calculated current distribution for this region of transition is given in Fig. 5(b) and is seen to approximate the measured distribution when $C_\lambda = 0.715$.

A comparison of the expressions used to obtain Figs. 5(a) and 5(b) shows that the amplitude of the two waves associated with the $T_0$ mode become more sharply attenuated as $C_\lambda$ is increased. This results in less total energy reaching the open end of the helix, and therefore, less energy reflected to create standing waves. The traveling waves associated with the $T_1$ mode begin to take on more and more importance in this region of transition, not only because of the increased attenuation encountered by the $T_0$ waves, but also because of the increase in the relative value of the constant amplitude of the $T_1$ waves compared to the amplitude of the $T_0$ waves at $x = 0$.

### C. The $T_1$ Mode Distribution

The $T_1$ or beam mode of radiation from the circular helix may be said to occur in the range of frequencies where $C_\lambda$ takes on the values 0.8 to 1.3. Figs. 3 and 4 show measured current distributions obtained in this region. A study of these data reveals that although the attenuated waves associated with the $T_0$ mode are always in evidence they attenuate rapidly so that the $T_1$ waves predominate a few turns from the feed point. Most of the energy which reaches the open end of the helix is reflected in the form of a $T_0$ wave, but this also attenuates rapidly leaving only the $F_1^+$ wave to predominate over most of the helix. This results in only a small reflected wave reaching the feed end of the helix and accounts for the relatively constant input impedance as a function of frequency. Fig. 5(c) shows a calculated current distribution which should be compared with the measured distributions when $C_\lambda = 1.01$ (see Fig. 4(a)).

The amplitude characteristic of the current distribution in the $T_1$ mode region can best be explained by observing a trend which may be seen in Figs. 3 and 4. When $C_\lambda = 0.691$ (see Fig. 3(c)), there are two minima which occur at $x = 53$ and 123 cm. As $C_\lambda$ takes on higher values, the minimum, which occurs in the vicinity of $x = 120$ cm, first becomes more pronounced and then disappears when $C_\lambda$ has values greater than 1.0. The minimum in the vicinity of $x = 50$ cm goes through the same variation and happens to be very pronounced in the distributions which occur in the range of frequencies where the $T_1$ mode is said to predominate. Similar minima occur again in the distributions when $C_\lambda$ is greater than 1.4.

These minima can be interpreted as the points where the $F_0^+$ and $F_1^+$ waves interfere since the two waves are traveling in the same direction with different phase velocities. The depths of the minima are controlled by the relative magnitudes of the two waves when they are in phase opposition. The increase in attenuation which the $T_0$ mode currents encounter, as $C_\lambda$ is increased, causes the minima first to become more pronounced as the initially larger $F_0^+$ wave attenuates to a value almost equal to the smaller, but constant $F_1^+$ wave. Then as the attenuation of the $F_0^+$ wave becomes large, the magnitude of this wave becomes so small compared to the $F_1^+$ wave at a point of interference that the minima cannot be observed.

The position of these minima, created by the outward traveling waves, is sometimes obscured and other times accentuated by the presence of reflected energy from the open end of the helix. The reflected waves also propagate in at least two modes. When $C_\lambda$ has values between 1.0 and 1.3, the SWR created by the reflected wave varies in a beatlike fashion. The points of small SWR can be interpreted as points of interference between the $F_0^-$ and $F_1^-$ waves.

Plots are presented in Fig. 6 of the component traveling waves which combine to produce the distribution shown in Fig. 5(c). These plots show how the $F_0^+$ and $F_1^+$ waves add to give the total outward traveling current distribution as well as the manner in which the reflected $F_0^-$ and $F_1^-$ waves combine to give the total reflected current distribution. Fig. 6 reveals how the pronounced minimum in the amplitude characteristic is obtained by the interference of the $F_0^+$ and $F_1^+$ waves. In addition, one observes in the absence of any reflected waves that there is a very rapid phase advance in the vicinity of the point of maximum interference ($x \cong 1.25\lambda_0$). However, since the outward traveling currents are at such a low level at this point, the reflected energy predominates and causes a reversal of the direction of energy flow over this short segment of the helix. Therefore, the total phase characteristic encounters a dip. It is observed in Figs. 3 and 4 that this effect is encountered many times in measurements; however, occasionally the reflected energy is not large enough to create the phase reversal, and the inherent phase jump is measured. (See Figs. 4(b), 4(c)).

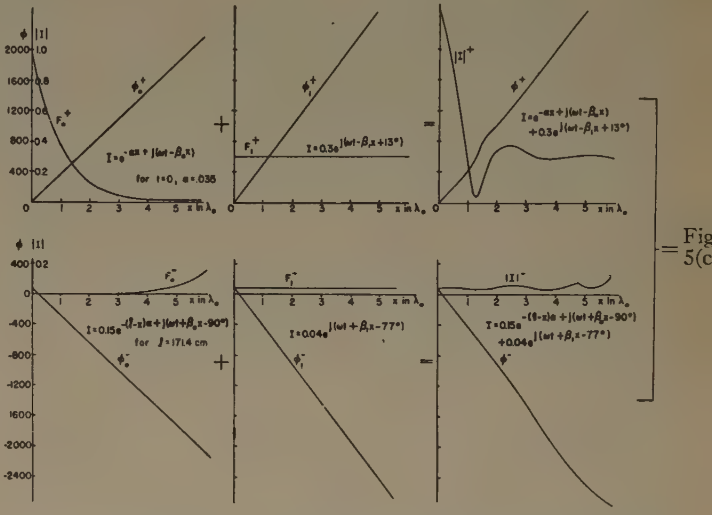

Fig. 6—Component traveling waves and the manner in which they add to produce a $T_1$ mode current distribution. (a) Outward traveling waves. (b) Reflected traveling waves.

### D. The Transition from the $T_1$ to the $T_2$ Mode

The effect of the $T_2$ mode currents upon the total current distribution may be observed in Fig. 4(e), 4(f), when $C_\lambda = 1.36$ and 1.50. The transition phenomena observed in this range of values for $C_\lambda$ is quite similar to those observed in the transition region between the $T_0$ and $T_1$ modes discussed above; therefore, only a few general observations will be made.

The presence of a $T_2$ mode on the helix has been detected by earlier investigators.[5,10] However, the radiation pattern of the helix in the range of frequencies where the presence of this mode has been detected[10] has broken up into a large number of lobes. Therefore, the investigation of the helix in this range of frequencies has not been stressed.

A calculated current distribution for the transition region from the $T_1$ to $T_2$ modes is illustrated in Fig. 5(d), and should be compared with the measured distributions shown in Figs. 4(d), 4(e), when $C_\lambda = 1.26$ and 1.36.

In the insert of Fig. 4(f), one will observe that the $F_1^+$ wave has an exponential attenuation. When $C_\lambda$ is made larger than 1.4, the $T_1$ mode also develops an

[10] C. K. Bagby, Jr., "A theoretical investigation of electromagnetic wave propagation on the helical beam antenna," an unpublished thesis for the degree M.Sc., The Ohio State University, 1948.

exponential attenuation similar to that of the $T_0$ mode discussed above.

## IV. Phase Velocity

It is apparent from the above analysis that there are always two modes of propagation present on the helix in the range of frequencies where a detailed investigation has been made. This means that there is no single phase velocity associated with the helix, except in certain regions of the helix at certain frequencies. At any given frequency there are always traveling waves which can be associated with two or more different modes, each of which has its own characteristic phase velocity. In analyzing the measured current distributions of Figs. 3 and 4 it is possible to distinguish the different modes present and their relative phase velocities. The results of this analysis may be seen in Fig. 7, where the $p$ versus $C_\lambda$ curves for the individual modes are given.

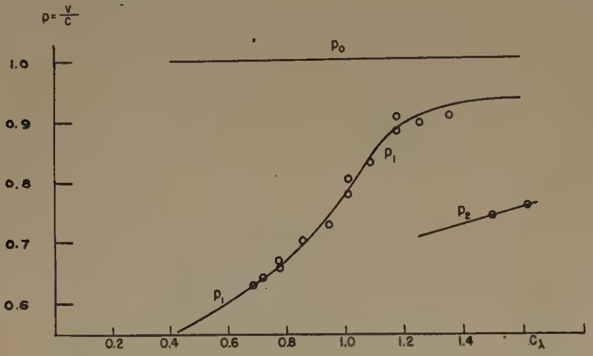

Fig. 7—$p$ versus $C_\lambda$ curves for the $T_0$, $T_1$, and $T_2$ modes on the helix.

It has been shown by Pierce,[11] and by Chu and Jackson,[12] that propagation along the helix in the $T_0$ mode has a characteristic phase velocity greater than the velocity of light. This velocity decreases as $C_\lambda$ increases and approaches closely to the velocity of light when $C_\lambda > 0.7$. Therefore, the relative phase velocity $p_0$ of the $T_0$ mode is assumed to be 1.0 at all larger values of $C_\lambda$. The maximum error introduced by this assumption occurs for small values of $C_\lambda$, and amounts to about 2 per cent when $C_\lambda = 0.7$.

In Fig. 7, one observes that $p_1$ and $p_2$ are functions of frequency and are always less than 1.0 in the range of the measurements. The quantities $p_1$ and $p_2$ are the phase velocities of the traveling waves associated with the $T_1$ and $T_2$ modes, respectively, relative to the velocity of light. The relative phase velocities assigned to each of the traveling waves in the inserts of Figs. 3 and 4 were obtained from Fig. 7. The information available in the inserts of Figs. 3 and 4, therefore, should enable one to calculate, to a good approximation, the far field-radiation pattern of the helix. The radiation pattern of the helix operating in the beam mode has been calculated by Kraus,[2,5] assuming that the current distribution is essentially a single unattenuated outward-traveling wave. This is a satisfactory approximation for a long helix. Kraus uses a phase velocity measured in the central portion of a seven-turn helix and good agreement is obtained between calculated and measured patterns.

The inserts of Figs. 3 and 4 show that the current distribution on the helix when operating in the $T_1$ mode is essentially the $F_1^+$ wave. Therefore, in the $T_1$ mode the $p_1$ versus $C_\lambda$ curve should agree fairly closely with a measured $p$ versus $C_\lambda$ curve obtained by Kraus.[13] The two curves are plotted in Fig. 8 for comparison. The maximum directivity line also shown in Fig. 8 was obtained from an expression derived by Kraus.[5] This expression gives the value $p$ should have for the radiation from an array of helical turns, considered as point sources, to add up along the axis of the helix in such a manner as to produce maximum directivity in the axial direction. In Fig. 8, it is observed that the $p_1$ curve, the maximum directivity curve, and Kraus' curve are all in close agreement within the region of the $T_1$ mode.

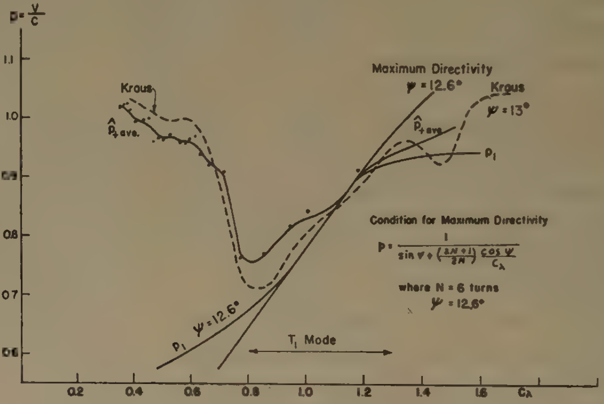

Fig. 8—$p$ curves for the uniform circular helix.

Since the task of calculating the radiation pattern for the helix is laborious at best, it would be desirable to reduce to at least two the number of traveling waves considered, even outside the region of the $T_1$ mode. In the $T_1$ mode, the $F_1^+$ wave alone may be used to calculate the patterns as has been done by Kraus. However, at both ends of the $T_1$ mode region and beyond, the traveling waves associated with the other modes cannot be neglected.

To facilitate calculations, it is suggested that all of the energy which flows in the positive $x$ direction be considered as one traveling wave, which has a phase velocity that is weighted in terms of the relative magnitudes of the component traveling waves which it is approximating. The energy which flows in the negative $x$ direction can be treated in the same manner. The

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[11] J. R. Pierce, "Theory of the beam-type traveling wave tube," Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.

[12] L. J. Chu and D. Jackson, "Field theory of traveling wave tubes," Proc. I.R.E., vol. 36, pp. 853–863; July, 1948.

[13] J. D. Kraus, "Antennas," McGraw-Hill Book Co., Inc., New York N. Y., in press; see Figs. 7–19.

generalized phase velocity, defined above, is such a weighted phase velocity. In the inserts of Figs. 3 and 4, one will notice curves which are labeled $p_+$. These curves are plots of the generalized phase velocity for the energy traveling in the positive $x$ direction only. These curves were obtained from calculations made on the associated measured current distribution. In drawing the $p_+$ curves, the regions where the associated $\phi$ versus $x$ curve dips were omitted for two reasons: first, the peculiar behavior in the region of the phase dips is almost always created by the reflected wave and, therefore, should not be allowed to affect $p_+$; second, the current in this region is always down in amplitude from the maximum by a factor of 20 db and the effect this small segment of the helix can have on the over-all radiation must be negligible.

The shape and trends shown by these $p_+$ curves are interesting. Each of the four regions discussed have $p_+$ curves which may be considered characteristic of that region. In the region of the $T_0$ mode, the $p_+$ curve is fairly constant and has a value between 0.95 and 1.0. The $p_+$ curve has an oscillatory nature in the regions of transition from one mode to another. The oscillations are similar to those observed when two waves are traveling at different velocities in the same direction. The center of oscillation, of course, shifts down as the $T_1$ mode currents tend to predominate. The $p_+$ curves in the $T_1$ mode region are characteristically high for low values of $x$ where the $T_0$ mode currents are still large. As the value of $x$ increases, the $p_+$ curve settles down to the value $p_1$ for most of the remaining length of the helix. The $p_+$ curves are often more sensitive indications of the presence of other modes than are the amplitude and phase curves themselves.

Each of the generalized $p_+$ curves was integrated graphically and the average value plotted in Fig. 8. The curve drawn through these points is labeled $p_{+\text{ave}}$. The values of $p$, obtained from this curve, can be used as a weighted phase velocity for a single traveling wave approximation of the energy traveling in the positive $x$ direction. It is observed that this curve follows roughly Kraus' curve in Fig. 8. Differences between these two curves should be expected since the $p_{+\text{ave}}$ curve is an average taken over the entire helix, whereas Kraus' curve was measured in the central region of a seven-turn helix.

## V. CHARGE AND CURRENT ON A LONG HELIX

The above analysis has been made almost entirely on the basis of data obtained from the helix shown in Fig. 2. As a check on this analysis, a second helix was constructed similar to the antenna of Fig. 2 but with 3.75 additional turns.

From the analysis made on the shorter helix, one should be able to predict the effect of extending the helix in length. For example, when $C_\lambda = 0.774$ (see Fig. 3(e)), the total current distribution is relatively free of any reflected energy when $x < 80$ cm. Extending the length of the helix at this frequency, therefore, should not alter the current distribution appreciably in the region $x < 80$ cm. However, by making the helix longer one would expect the amplitude curve in the vicinity of $80 \leqq x \leqq 160$ cm to be smoothed out since the attenuated reflections would occur at some point beyond $x = 172$ cm.

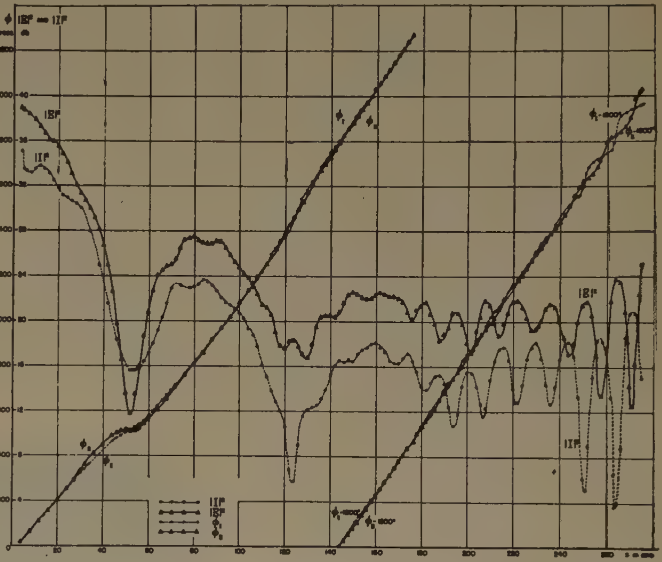


Fig. 9—Measured current and charge distribution on a 9.75-turn helix, pitch angle $\psi = 12.6°$, mean diameter $D = 8.61$ cm for $C_\lambda = 0.774$.

The current distribution shown in Fig. 9 was measured on the long helix, 9.75 turns, for $C_\lambda = 0.774$. Comparing Fig. 9 with Fig. 3(e) for the six-turn helix at the same value of $C_\lambda$, one finds that all of the predictions are fulfilled.

As a further check on the analysis, a charge or $E$ field measurement as a function of distance along the long helix was made and plotted in Fig. 9. Comparing the $E$ and $H$ field measurements shown in Fig. 9, one observes that for low values of $x$, where the outward traveling waves predominate, the $E$ and $H$ fields are in phase and have the same general shape. However, as $x$ takes on larger values, the effect of the reflected energy causes the $E$ and $H$ fields to approach a relative phase quadrature. That is, the two phase characteristics take on an interlaced appearance, a maximum $E$ field amplitude being found where a minimum $H$ field is observed, and vice versa.

## VI. ACKNOWLEDGMENT

The advice and encouragement of J. D. Kraus, and the assistance received from the staff of the Ohio State University Research Foundation Antenna Laboratory, are gratefully acknowledged.

# An Atmospherics Waveform Receiver*

WILLIAM J. KESSLER†, ASSOCIATE, IRE, AND SYDNEY E. SMITH†

A KNOWLEDGE of the instantaneous field-intensity variations, or waveforms, of atmospherics noise provides a useful starting point for studying the characteristics and mechanism of lightning discharges, as well as many radio propagational phenomena.

The present summary constitutes a brief description of a driven-sweep oscillograph of special design for the observation of atmospherics waveforms. A 15-foot vertical antenna coupled to a combination signal-power dual coaxial transmission line through a cascaded cathode-follower impedance transformer is provided to permit locating the antenna in a remote interference-free area.

The signal amplifier employs resistance-capacitance parallel-T filter sections exhibiting transmission nulls at one megacycle and at 60 cycles per second to eliminate interfering signals due to radio-broadcast transmitters and further reduce the interference from stray power-line fields. An 18-microsecond delay line permits delineation of the entire leading edge of the atmospherics waveform. A sweep control circuit is arranged to trigger the horizontal sweep generator, regardless of the polarity of the initial maximum of the atmospherics. Limiting of the display-rate to about eight waveforms per second prevents overlapping of the waveforms on the photographic record.

A marker generator permits the measurement of time intervals and provides a continuous check upon the stability and linearity of the sweep generator. The display of the markers, spaced either 50 or 100 microseconds, is unique in that they appear on a second sweep displayed 2 inches above, rather than superimposed upon the atmospherics waveforms. Confusion of the markers with the characteristic variations of the waveform is thus prevented. This display is accomplished with a cathode-ray tube containing a single electron gun by blocking the signal channel during the application of the marker sweep which is delayed approximately 1,500 microseconds after the end of the signal sweep.

The atmospherics waveform receiver was developed to implement a research program to investigate the possibilities of thunderstorm location through waveform analysis of the associated atmospherics. Further applications may be found in other geophysical areas, such as the mapping of subterranean strata with the aid of artificially injected earth shocks or the investigation of the lower ionospheric regions employing atmospherics as natural pulses. The application of the atmospherics waveform receiver for ionospherics research is of particular interest because of the information provided on the ionosphere in the low-frequency region down to 10 or 20 kc.

Fig. 1 is an atmospherics waveform observed during the night. Successive pulses which are approximate replicas of the parent pulse that have experienced multiple reflections between the lower regions of the *E* layer and the surface of the earth are clearly visible. Processing of the time-intervals between successive reflections observed in Fig. 1 through a graphical procedure based on a method of least squares yields a great circle distance of approximately 1,175 kilometers to the discharge source and a mean reflection height of about 90 kilometers.

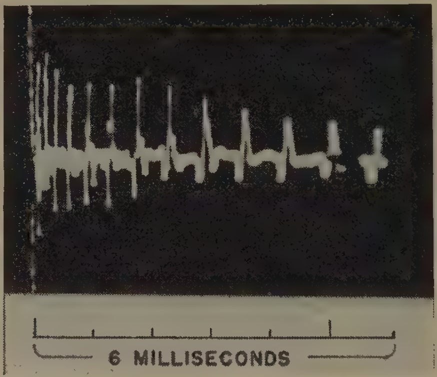
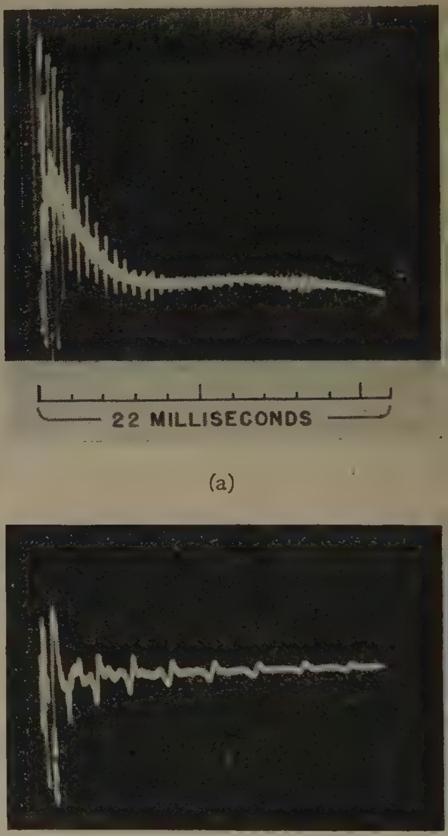


Fig. 1—Characteristic atmospherics waveform recorded during night.

Examples of multiple returns from lightning discharges associated with an overhead thunderstorm in Gainesville are shown in Fig. 2(a). Approximately 20 vertically-incident returns are shown which exhibit a delay of about 610 microseconds between consecutive returns corresponding to an effective reflection height of about 91 km. Fig. 2(b) shows another waveform recorded during similar conditions with an expanded time scale. The effects of energy absorption and the small degree of dispersion are evident.



(a)

(b)

Fig. 2—Multiple returns from overhead thunderstorm; (a) 22-millisecond sweep; (b) 5-millisecond sweep.

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* Decimal classification: R272.1. Original manuscript received by the Institute, May 10, 1950; abstract received, October 11, 1950. Presented, IRE National Convention, New York, N. Y., March 9, 1950.

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# On the Response of a Directive Antenna to Incoherent Radiation*

FREDERICK W. SCHOTT†, MEMBER, IRE

*Summary*—**The relative response of directive antennas is dependent on the mechanism whereby the energy is conveyed. This fact, which often seems to be neglected, is of particular importance when the energy arriving at a receiving antenna can be represented in terms of a spectrum of incoherent plane-wave components. Such may be true frequently in the extraoptical propagation of microwaves.**

**A study of an idealized situation is presented which compares the behavior of several pencil-beam antennas of varying aperture. A brief comparison with some recent observations is made.**

## INTRODUCTION

THE CONCEPT of antenna gain is a useful tool for computing the power received by an antenna when the incident radiation consists of a single, substantially plane, wave. However, when the incoming radiation is of a more complex nature as, for example, when due to an aggregate of scattering elements randomly located, the gain concept must be applied with some caution.

The purpose of this paper is to indicate the manner in which the power received by a directive antenna depends upon the distribution of the incident radiation and to illustrate the result with an application to a pencil-beam antenna.

## THE WEIGHTED ANTENNA GAIN

A directive antenna whose axis is chosen as the reference direction is shown in Fig. 1. A parabolic antenna is shown; however, the form of the antenna is immaterial at this point. It is supposed that the incident radiation consists of many plane-wave components, and it is assumed that the $n$th component is arriving from the direction $\theta_n$, $\phi_n$, where $\theta_n$ represents the colatitude and $\phi_n$ the azimuth. Because of the directive characteristics of the antenna, this wave produces a component of voltage of relative amplitude $a_n$ at the antenna output, i.e.,

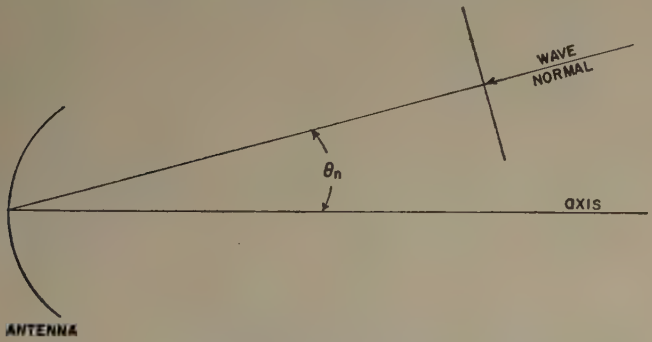

Fig. 1—Directive antenna and associated notation.

$$a_n = a(\theta_n, \phi_n).$$

The resultant amplitude (voltage) at the antenna output due to the many plane waves is then

$$E = \sum_{n=1}^{N} a_n e^{j\alpha_n}, \tag{1}$$

where $\alpha_n$ represents the phase of the $n$th component relative to some arbitrary reference and is assumed to be *random.*

The power associated with this resultant voltage (except for a constant of proportonality) is then

$$S = |E|^2 = \left|\sum_{n=1}^{N} a_n e^{j\alpha_n}\right|^2 \tag{2}$$

and the average power

$$\overline{S} = \sum_{n=1}^{N} a_n^2 \tag{3}$$

because of the random nature of the $\alpha$'s. If the plane-wave components are sufficiently numerous, the above summation over $n$ can be represented by a corresponding integration over the appropriate directions. Since the element of area is proportional to $\sin\theta d\theta d\phi$, the average is given by[1]

$$\overline{S} = C\int_0^{2\pi}\int_0^{\pi} a^2(\theta, \phi)\sin\theta d\theta d\phi, \tag{4}$$

where $C$ is a constant and $a(\theta, \phi)$ now represents the probability per unit area that the component of amplitude $a$ is associated with the direction $\theta$, $\phi$.

Because $a(\theta, \phi)$ is a measure of the relative amplitude at the antenna output, it represents the effect of antenna directivity as well as the directional character of the radiation itself. Hence

$$a^2(\theta, \phi) = a_1^2(\theta, \phi)G(\theta, \phi),$$

where $a_1(\theta, \phi)$ is the relative amplitude of the plane-wave component in the direction of polarization of the antenna arriving from the direction $\theta$, $\phi$, and $G(\theta, \phi)$ is the gain function[2] of the antenna in a homogeneous medium.

If an isotropic antenna were used, the average received power would be given by

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* Decimal classification: R125×R326.8. Original manuscript received by the Institute, May 9, 1950; revised manuscript received, October 2, 1950. Presented, IRE-URSI Symposium on Antennas and Propagation, San Diego, Calif., April 3, 1950.

† University of California, Los Angeles, Calif.

[1] See P. M. Morse, "Vibration and Sound," McGraw-Hill Book Co., New York, N. Y., p. 383; 1948.

[2] As defined, for example, by S. Silver, "Microwave Antenna Theory and Design," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 12, p. 3; 1949.

$$\overline{S}_I = C \int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) \sin \theta d\theta d\phi. \tag{5}$$

Thus, the power received by the directive antenna relative to that received by an isotropic antenna is

$$\frac{\overline{S}}{\overline{S}_I} = \frac{\int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) G(\theta, \phi) \sin \theta d\theta d\phi}{\int_0^{2\pi} \int_0^{\pi} a_1^2(\theta, \phi) \sin \theta d\theta d\phi}, \tag{6}$$

which might be termed the "weighted" gain. This rather general result must include, as a special case, that of radiation arriving from a single direction, $\theta_1$, $\phi_1$. In the latter case the directivity of the incident radiation may be represented by the delta function, i.e., $a_1(\theta, \phi) = \delta(\theta - \theta_1, \phi - \phi_1)$, giving

$$\frac{\overline{S}}{\overline{S}_I} = G(\theta_1, \phi_1),$$

as required.

It should be noted that the result using (6) is determined by the method used to illuminate the scattering elements as well as their own directivity characteristics, since this determines $a_1(\theta, \phi)$. An analysis by Booker and Gordon,[3] based on the work of Pekeris,[4] has given explicit results for certain cases of forward scattering due to atmospheric inhomogeneities.

## EXAMPLES

As an illustration of the use of (6), the following case will be considered: It will be assumed that the antenna is viewing the "center" of a scattering region such that $a_1(\theta, \phi) = a_1(\theta)$ only, and furthermore that the incident plane-wave components arrive uniformly from all directions within the cone semiangle $\theta_0$, and that no radiation is arriving from outside this region. This is indicated in Fig. 2. In addition, it will be assumed that

$$G(\theta, \phi) = G(\theta)$$

only. Thus (6) becomes

$$\frac{\overline{S}}{\overline{S}_I} = \frac{\int_0^{\theta_0} G(\theta) \sin \theta d\theta}{\int_0^{\theta_0} \sin \theta d\theta}. \tag{7}$$

The assumptions leading to (7) would be, in general quite severe; however, there are instances where highly directive antennas are used in which such restricted results may be of value.

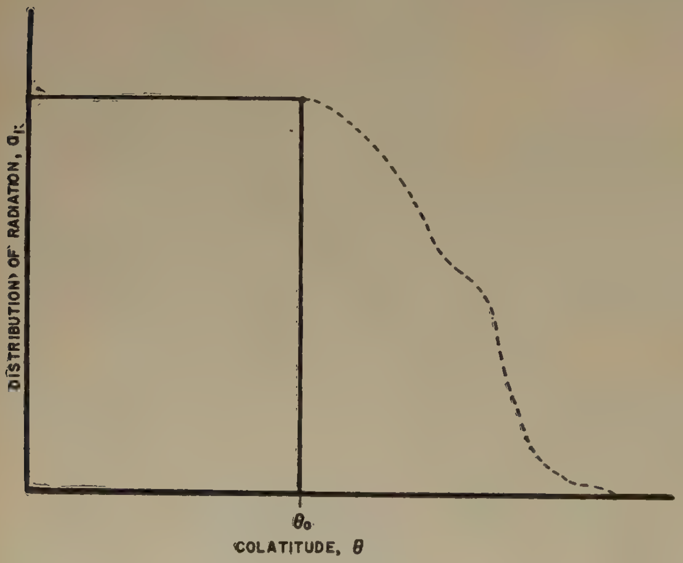

Fig. 2—A theoretical distribution of incident radiation.

To obtain specific results, four parabolic antennas having apertures of 12, 18, 30, and 48 inches, respectively, have been studied, using (7). For each antenna the gain function in the $E$ and $H$ planes was measured at 9,375 mc and the average used as $G(\theta)$. This procedure is appropriate[5] inasmuch as the significant portion of the pattern is nearly identical in the two planes.

The results, expressed as a function of $\theta_0$, are shown in Fig. 3.

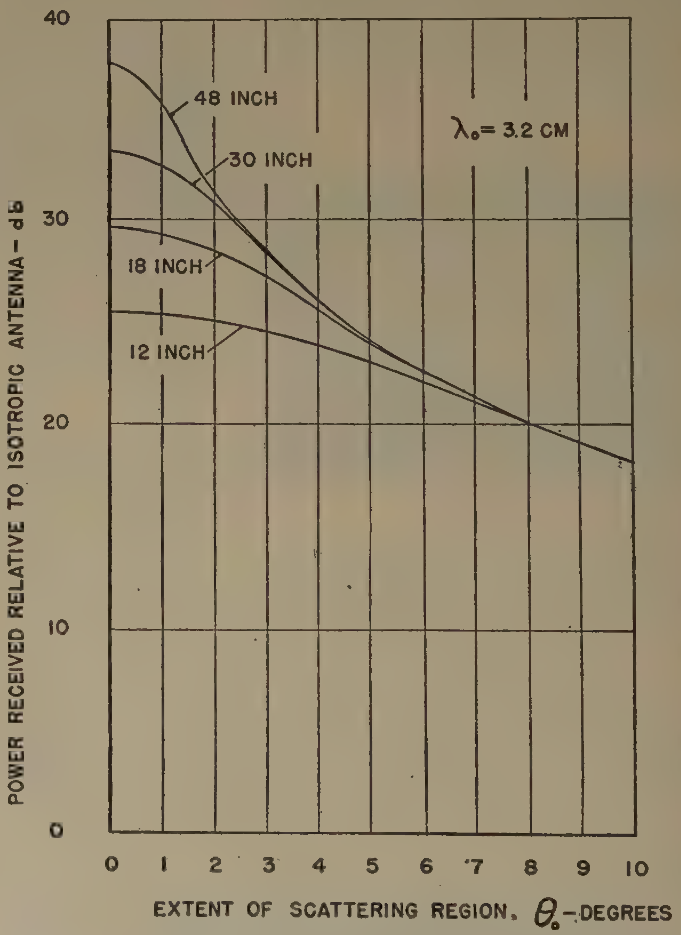

Fig. 3—The response of parabolic antennas to scattered radiation.

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[3] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," PROC. I.R.E., vol. 38, pp. 401–413; April, 1950.

[4] C. L. Pekeris, "Note on scattering in an inhomogeneous medium," *Phys. Rev.*, vol. 71, p. 268; February, 1947.

[5] See page 413 of footnote reference 2.

It may be seen that if the radiation is arriving from a region which is sufficiently large, i.e., if $\theta_0$ is large relative to the beam width, the received power is independent of the antenna aperture. Furthermore, in a case such as this, where the minor lobe pattern is insignificant, the incident radiation could vary in any manner outside the main portion of the beam, such as indicated by the dotted curve of Fig. 2, for example, without invalidating the above observation.

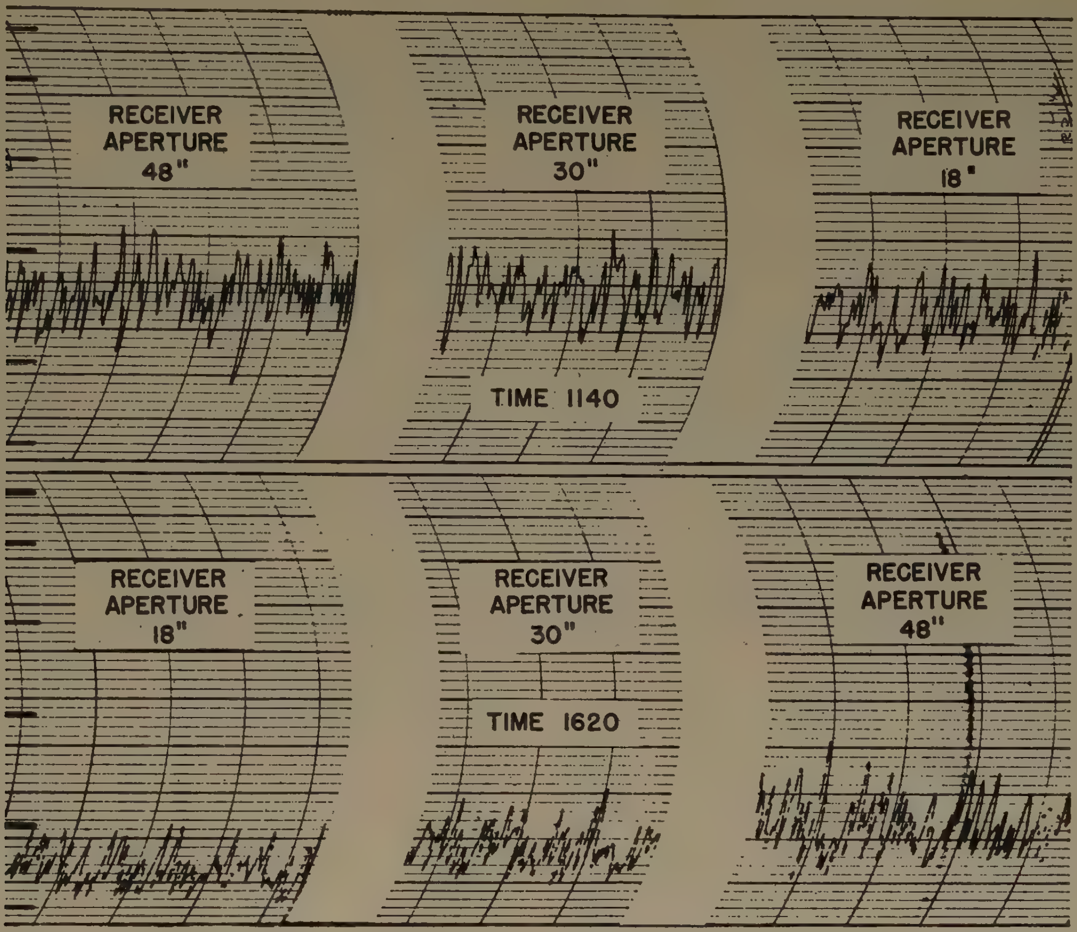

Fig. 4—Relative response of directive antennas. (March 6, 1950; 9,340 mc; transmitter aperture, 18 inches; transmitter and receiver cab elevation, 0 foot; separation, 46.2 miles.)

Microwave propagation experiments in the Arizona Desert have indicated that in many cases scattering may have been the dominant mechanism involved, and recent studies have been made to learn more of the nature of these scattered fields. Although a complete analysis of these data will follow in a later paper, two preliminary observations may be noted:

1. Only during the evening, under conditions of a strong temperature inversion, or during the day when high in the diffracted region, was the average received power proportional to the gain of the receiving antenna. At these times the fields did not fluctuate greatly.

2. During the daytime with low antenna heights, when rapid scintillations of large amplitude were noted, the average received power would usually increase when a higher gain antenna was used, but never in proportion to the gain increase. In a typical situation, changing the aperture from 18 to 30 and from 30 to 48 inches corresponded to measured gain increments of 3.7 and 4.5 db. whereas the observed power increased in steps of 2.1 and 2.4 db, respectively. This is illustrated in the lower portion of Fig. 4. The upper data represent a less common but more striking example of this phenomenon. (The markers at the left indicate 10-db intervals.)

## CONCLUSIONS

There are many situations where an arriving wave can be represented only by a spatial spectrum of essentially incoherent plane wave components. This seems to be true often in the extraoptical propagation of microwaves,[6] and it is felt to be of significance in the optical range in the presence of appreciable rain attenuation.

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[6] Recent experimental results presented by A. W. Straiton and D. F. Metcalf at the IRE-URSI Symposium on Antennas and Propagation held in San Diego, during April 3–5, 1950, suggest that extra-optical propagation at FM frequencies may also be due to atmospheric scattering.

When it is desired to predict the manner in which a particular receiving antenna will perform under such circumstances, considerations of the type indicated may be appropriate.

In this connection it must be emphasized that the frequently used engineering procedure for expressing the power associated with an incident field may be misleading and erroneous. It is quite common to find fields expressed in decibels relative to a particular microvolt-per-meter level.[7] This is accomplished by the use of a calibration which is appropriate when the wave front is essentially plane; however, this same calibration is then used *regardless of whether or not the radiation is known to be of this type.*

If the distribution of the incident radiation is unknown, one can only state the power received by the particular antenna configuration used.

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[7] E. W. Allen, W. C. Boese, and H. Fine, "Summary of Tropospheric Propagation Measurements and the Development of Empirical VHF Propagation Charts (Revised)," Federal Communications Commission, T.I.D. Report No. 2.4.6.; May 26, 1949.

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# Optimum Operation of Echo Boxes*

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***Summary*—The parameters determining ring time for an echo box in a radar system are examined, and the conditions for optimum ring time established. Two specific arrangements, namely, the echo box coupled directly to the guide and the echo box coupled through a directional coupler, are treated in detail.**

**Representative curves of ring time as functions of various parameters, and curves of power extracted by the echo box are included. It will be seen that simply increasing the loaded *Q* of a cavity will not increase the ringing time past a certain maximum, and that there is an optimum value of loaded *Q* which is a function of the radar power, receiver sensitivity, pulse length, and the method of coupling the cavity to the transmitter.**

## I. INTRODUCTION

THE USE OF echo boxes in checking over-all radar system performance is well known.[1,2] Although the design of the cavity per se has been treated in considerable detail,[1,3] and the general operation of echo boxes and other resonant cavities discussed,[2,4,5] the full significance of the factors involved in obtaining maximum ring time from a resonant cavity used in conjunction with a particular radar is not too readily apparent. It is the purpose of this paper to consider in detail these factors. Two methods of coupling the echo box to the waveguide are considered. In one, the echo box cavity is mounted directly on the side of the guide. In the other, it is mounted on the end of a secondary guide, which in turn is coupled to the main guide through a directional coupler.

The behavior of the echo box is calculated for the following assumed conditions. The transmitted pulse is assumed to be of constant amplitude and frequency throughout its duration; i.e., the rise and decay times are assumed negligible, compared to the pulse duration, and the pulse is assumed to be free from frequency modulation. The transmission line is assumed to be terminated in its characteristic impedance, both at the antenna and at the transmitter-receiver. If a particular system is known to depart significantly from the above assumed conditions, corrections may be made in the calculations, but the same general principles should apply. For example, if the transmitter is known to be frequency modulated, the amount of energy available to an echo box tuned to a specific frequency will be reduced, and so forth.

As in the case of any other measuring device, it is desirable that the effect of the measuring device on the performance of the system be negligible. Hence, it is advisable that the introduction of the echo box into the system have no appreciable effect, either on the impedance seen by transmitter or on the power delivered to the antenna. This requirement may preclude the adjustment for maximum ring time. The limitations are discussed in some detail below.

## II. LIST OF SYMBOLS

$R$ = equivalent cavity resistance

$C$ = equivalent cavity capacitance

$L$ = equivalent cavity inductance

$Z_0$ = characteristic impedance of transmission lines

$e$ = voltage across cavity

$e_{\min}$ = minimum detectable value of $e$

$E$ = transmitter voltage across matched line

$R_g$ = generator impedance in equivalent circuit

$E_g$ = generator voltage in equivalent circuit

$e_r$ = voltage across receiver

$K$ = voltage coupling factor of directional coupler

$Q_L$ = loaded $Q$ of cavity

$Q_0$ = unloaded $Q$ of cavity

$a$ = length of exciting pulse

$\omega_0$ = resonant angular frequency

$S_{-1}$ = unit step function

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† Raytheon Manufacturing Company, Newton, Mass.

[1] I. G. Wilson, C. W. Schramm, and J. P. Kinzer, "High $Q$ resonant cavities for microwave testing," *Bell Sys. Tech. Jour.*, vol. 25, pp. 408–434; July, 1946.

[2] E. I. Green, H. J. Fisher, and J. G. Ferguson, "Techniques and facilities for microwave radar testing," *Bell Sys. Tech. Jour.*, vol. 25, pp. 435–482; July, 1946.

[3] J. P. Kinzer and I. G. Wilson, "Some results on cylindrical cavity resonators," *Bell Sys. Tech. Jour.*, vol. 26, pp. 410–445; July, 1947.

[4] C. G. Montgomery, "Technique of Microwave Measurements," vol. 11, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., p. 341; 1947.

[5] "Notes on Lectures of W. W. Hansen," Radiation Laboratory, (limited classified printing), chap. 12; 1943.

$p$ = complex variable of Laplace transform
$t$ = time
$j=\sqrt{-1}$
$T$ = ring time of cavity measured from end of pulse
$F$ = voltage "dynamic range" constant of transmitter-receiver combination
$M$ = ratio of power taken by echo box to power delivered to load
$A$ = ratio of average power into echo box, short pulse, to average power, long pulse.

## III. BASIC COUPLING SYSTEMS

Figs. 1 and 2 are schematic representations of the two basic types of coupling systems herein considered. Fig. 1 shows a cavity coupled directly to the line, and Fig. 2 shows a directional-coupler type of connection. The lumped circuit equivalent of both of these two circuits is shown in Fig. 3.

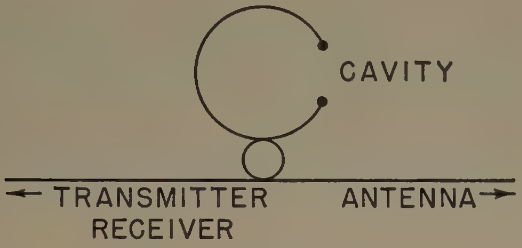

Fig. 1—Echo box coupled directly on line.

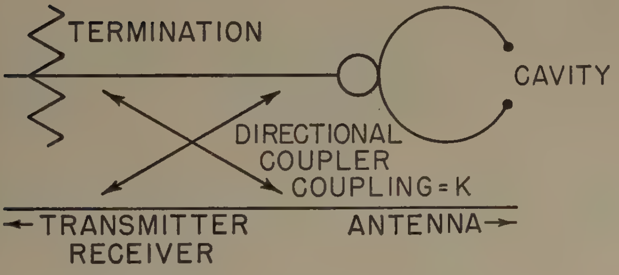

Fig. 2—Echo-box system using directional coupler.

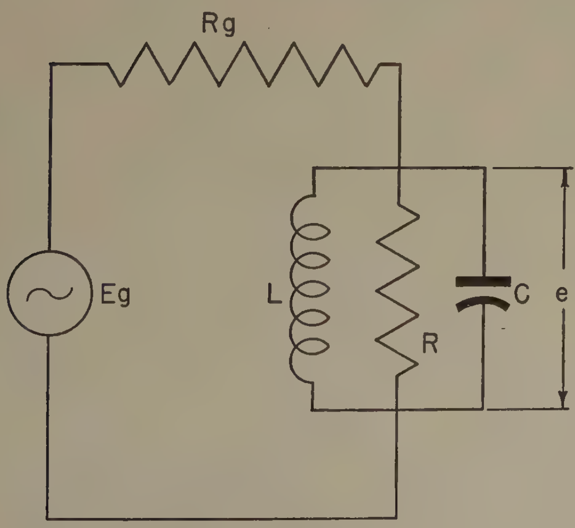

Fig. 3—General equivalent circuit.

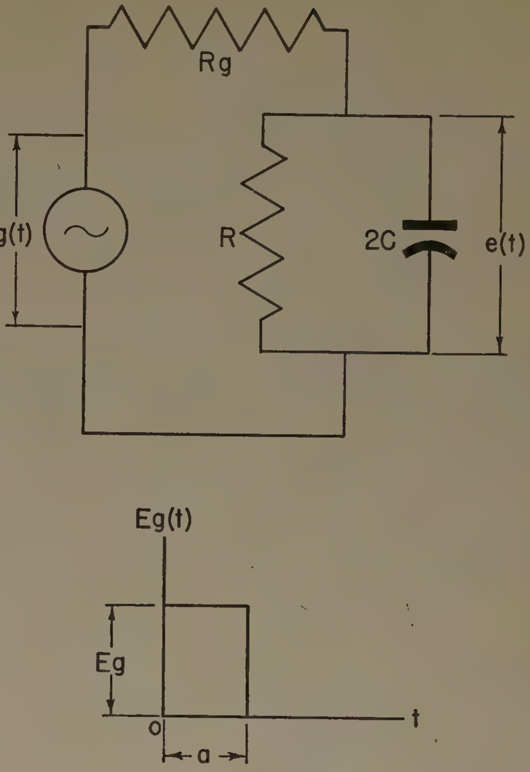

Fig. 4—Low-pass analog.

The transmitter can be treated as an open-circuit voltage $2E$ in series with an internal impedance $Z_0$; the voltage across the matched line to the antenna is then $E$, and the equivalent generator voltage in the auxiliary line of Fig. 2 is $E/K$. If the voltage across the cavity at any instant is $e$, the voltage across the receiver, during the decay of the "echo" is $e_r=e/2$ in Fig. 1 and $e_r=e/K$ in Fig. 2.[6]

The following table gives the equivalent circuit values for each case:

TABLE I

| Equivalent Circuit Fig. 3 | Cavity on Line Fig. 1 | Directional Coupler Fig. 2 |
|---|---|---|
| $R_g$ | $2Z_0$ | $Z_0$ |
| $E_g$ | $2E$ | $E/K$ |
| $e$ | $2e_r$ | $Ke_r$ |

The basic relations derived from elementary resonant circuit theory, between the parameters of the circuit shown in Fig. 3 and the resonant angular frequency $\omega_0$, loaded $\phi_L$, and unloaded $\phi_0$ of the cavity are

$$Q_0 = R\sqrt{\frac{C}{L}} \tag{1}$$

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[6] If the echo box is coupled to the radar system by means of an auxiliary antenna, instead of being coupled to the transmission line, the analysis can be made the same as for the directional coupler case by letting the equivalent generator voltage in the transmission line from the auxiliary antenna to the echo box equal $E/K$.

$$Q_L = \frac{R_g R}{R_g + R}\sqrt{\frac{C}{L}} \qquad (2)$$

$$\omega_0 = \frac{1}{\sqrt{LC}} \qquad (3)$$

$$\frac{R}{R + R_g} = 1 - \frac{Q_L}{Q_0} \qquad (4)$$

$$\frac{R_g + R}{2RR_gC} = \frac{\omega_0}{2Q_L}. \qquad (5)$$

In order to analyze the response of the above circuit to a pulse-modulated sine wave, we make a conventional band-pass to low-pass transformation to arrive at the circuit of Fig. 4.[7] This transformation is permissible because of the small cavity bandwidth compared to its center frequency and the long pulse length compared to the period of the radio-frequency carrier. This circuit is now analyzed for its response to a rectangular pulse.

## IV. Calculation of Ringing Time

This analysis is performed by conventional Laplace transform methods. The transform of a rectangular pulse of width $a$ and height $E_g$, using the "real translation" theorem, is given as

$$E_g(p) = \frac{E_g}{p}(1 - \epsilon^{-ap}). \qquad (6)$$

Standard network theory yields as the transform of the output voltage

$$e(p) = \frac{\frac{E_g}{2R_gC}(1 - \epsilon^{-ap})}{p\left(p + \frac{R + R_g}{2RR_gC}\right)}. \qquad (7)$$

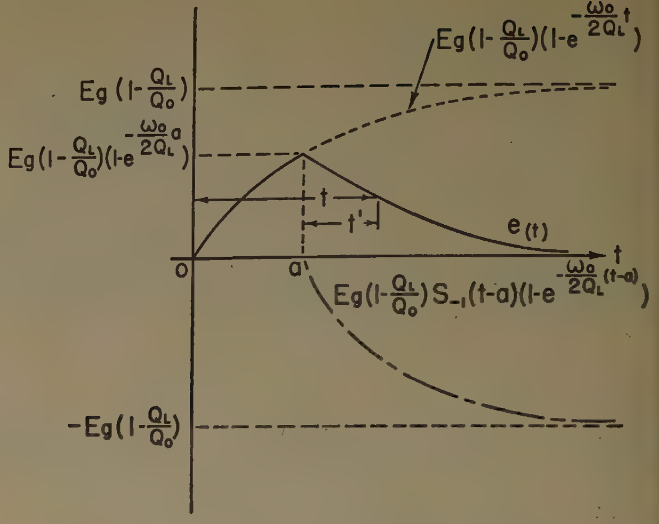
The inverse transform is found from tables of Laplace transforms, and (4) and (5) enable us to write

$$e(t) = E_g\left[1 - \frac{Q_L}{Q_0}\right][1 - \epsilon^{-\omega_0/2Q_L}$$
$$- S_{-1}(t - a)(1 - \epsilon^{-\omega_0/2Q_L(t-a)}]. \qquad (8)$$

Fig. 5 is a graphical picture of (8) and shows the effects of pulse length and coupling on the maximum amplitude of the energy in the cavity and on the ring time. The factor $S_{-1}(t-a)$ becomes equal to one for $t>a$. We define $t'=t-a$ and ring time $T$ as the value of $t'$ when $e(t')$ is the minimum detectable, $e_{\min}$. Equation (8) is then solved for $T$.

$$T = \frac{2Q_L}{\omega_0}\ln\left[F\left(1 - \frac{Q_L}{Q_0}\right)(1 - \epsilon^{-(\omega_0/2Q_L)a})\right], \qquad (9)$$

where $F = E_g/e_{\min}$.



$Eg(1-\frac{Q_L}{Q_0})(1-e^{-\frac{\omega_0}{2Q_L}t})$
$Eg(1-\frac{Q_L}{Q_0})$
$Eg(1-\frac{Q_L}{Q_0})(1-e^{-\frac{\omega_0}{2Q_L}a})$
t
t'
e (t)
t
0
a
$Eg(1-\frac{Q_L}{Q_0})S_{-1}(t-a)(1-e^{-\frac{\omega_0}{2Q_L}(t-a)})$
$-Eg(1-\frac{Q_L}{Q_0})$

Fig. 5—Graphical interpretation of (8).

Equation (9) should be differentiated with respect to $Q_L/Q_0$ and the derivative set equal to zero to determine the optimum ratio, but this is an awkward procedure because of the unwieldy transcendental equations. Hence families of curves have been plotted using values of the parameters representative of a wide range of typical radar systems. These curves are shown in Figs. 6 through 8.

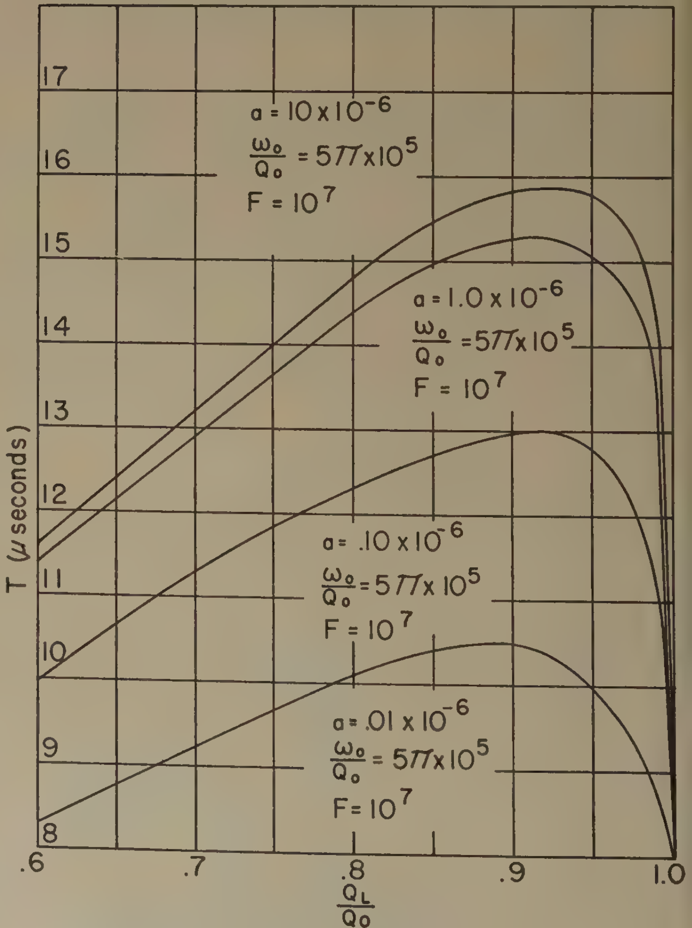


Fig. 6—Ring time versus $Q_L/Q_0$ for different pulse lengths.

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[7] Ernst A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. II; 1935.

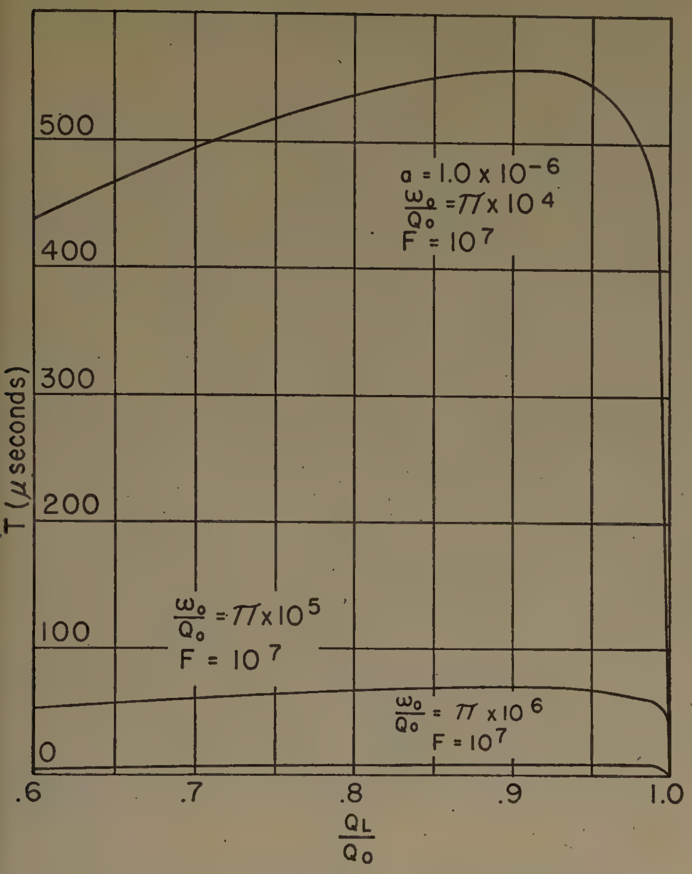

Fig. 7—Ring time versus $Q_L/Q_0$ for various unloaded $Q$'s.

We also note that for all system parameters constant, and for a constant ratio of loaded to unloaded $Q$, the loaded $Q$ is proportional to frequency for a given ring time.

Fig. 6, which is a plot of ring time versus $Q_L/Q_0$ for several different pulse lengths, clearly shows that there is an optimum loaded $Q$ which will yield maximum ring time. The ratio of $Q_L/Q_0$ is about 0.9 for a 0.01-microsecond pulse and gets progressively higher as the pulses get longer. This is to be expected physically since the shorter pulses do not give a high-$Q$ cavity sufficient time to store its maximum energy. It is important to note that if the cavity loaded $Q$ is too low, the loss in ring time will not be serious, but if it is too high the ring time will degenerate very rapidly.

Fig. 7 is similar to Fig. 6 except that all these curves are for the same pulse length (1 microsecond) but different ratios of frequency to unloaded $Q$. It is apparent from these curves that the ringing time is very sensitive to unloaded $Q$ and that every effort should be made to obtain as high an *unloaded* $Q$ as possible.

Fig. 8 has as a parameter $F$, the dynamic range in voltage between the transmitted signal and minimum discernible received signal. These curves show the optimum $Q_L/Q_0$ decreasing as $F$ decreases. A general conclusion to be drawn from these curves, along with those of Fig. 6, is that low-power short-pulse radars demand a lower optimum loaded $Q$ where the higher-power long-pulse systems can use a higher loaded $Q$ to achieve maximum ring time.

If a directional coupler is used to couple the cavity, then the factor $K^2$ appears in $F$ in accordance with Table I. Otherwise the computation for ring time with or without a directional coupler is identical.

The above curves cover the range of parameters normally encountered in microwave radars.

## V. Power Loss in Echo Box

For the case of the echo box coupled through a directional coupler, the power extracted from the main guide is proportional to the square of the voltage coefficient of coupling of the directional coupler

$$M = \frac{P_{\text{extracted}}}{P_{\text{load}}} = K^2. \qquad (10)$$

Thus, for example, if $K$ is less than 0.1, the maximum power extracted by the measuring system will be less than one per cent of the total power.

For the case of the cavity mounted directly on the guide, the average power extracted during the transmitted pulse is determined by the coupling and also by the pulse length.

For very long pulses, a steady-state condition may be assumed, the energy *stored* in the echo box may be neg-

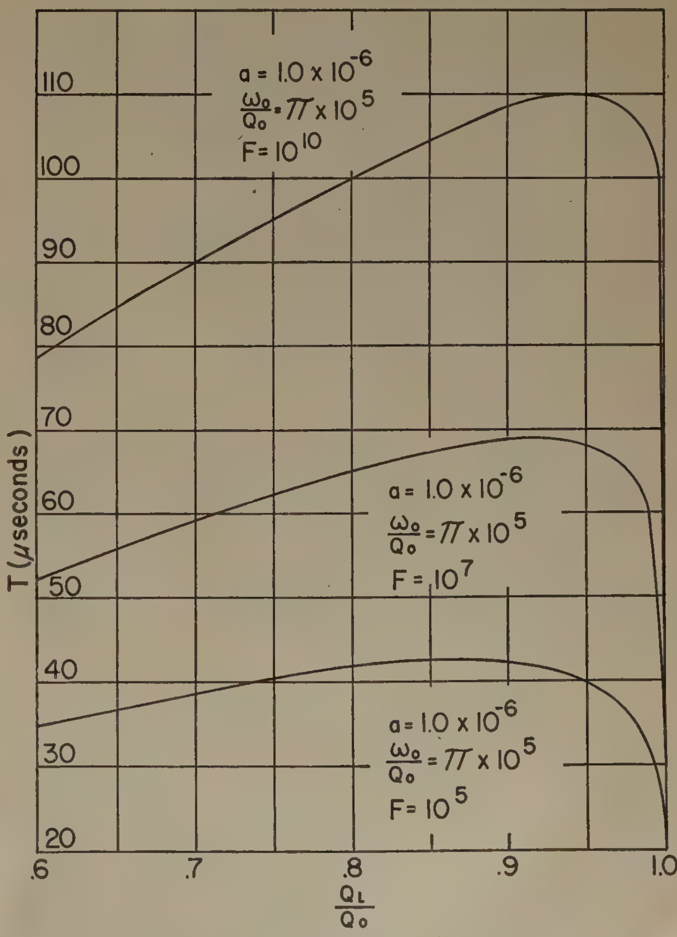

Fig. 8—Ring time versus $Q_L/Q_0$ for various dynamic ranges.

lected in comparison with the energy *dissipated* therein, and the ratio $M$ of power into the echo box to power into the load is accordingly given by

$$M = \frac{2R}{R_g}, \tag{11}$$

wherein $R$ and $R_g$ are shown in Fig. 3. Relating $R$ and $R_g$ to $Q_L$ and $Q_0$ by (4), (11) becomes

$$M = 2\left(\frac{Q_0}{Q_L} - 1\right). \tag{12}$$

This is plotted in Fig. 9 as the curve for infinite pulse length ($\omega_0 a/2Q_L = \infty$). For short pulses, the average power into the echo box will be less than indicated in (12) because of the effect of the initial transient condition. The ratio of power extracted during a short pulse to power delivered to the load can be approximated by multiplying the above factor $M$ by $A$, where $A$ is the ratio of average power into the echo box, for the short pulse, to the power into the echo box for the steady-state condition.

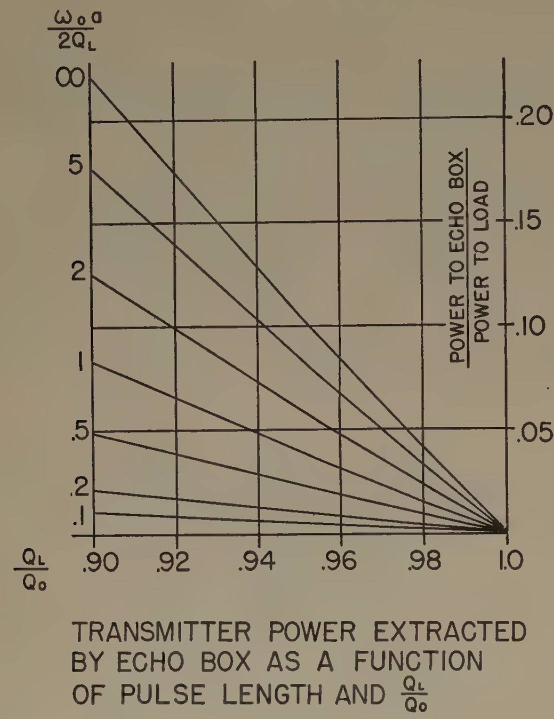

Fig. 9—Ratio of cavity power to load power versus $Q_L/Q_0$.

The energy delivered to the echo box in time $a$, in the steady-state condition, would be

$$\frac{\left[E_g\left(1 - \frac{Q_L}{Q_0}\right)\right]^2 a}{R}.$$

The energy dissipated in the echo box, during a pulse of length $a$, is

$$\frac{1}{R}\int_0^a e^2 dt.$$

The energy *stored* in the echo box at the end of the pulse is

$$e^2(a)\frac{Q_0}{\omega_0 R}.$$

Hence

$$A = \frac{\frac{1}{R}\int_0^a e^2 dt + e^2(a)\frac{Q_0}{\omega_0 R}}{\left[E_g\left(1 - \frac{Q_L}{Q_0}\right)\right]^2 a/R}. \tag{13}$$

As

$$e = E_g\left(1 - \frac{Q_L}{Q_0}\right)(1 - \epsilon^{-\omega_0 t/2Q_L}), \tag{14}$$

it can be shown that $A$ is given approximately by the equation

$$A = 1 - \frac{2Q_L}{\omega_0 a}(1 - \epsilon^{-\omega_0 a/2Q_L}). \tag{15}$$

Multiplying the curve given in Fig. 9 for very long pulses by this factor gives the other curves shown in that figure. These curves would indicate that radar systems using very short pulses can tolerate much tighter coupling to an echo box than would normally be used.

## VI. Conclusions

The dependence of the optimum $Q$ and maximum ring time of an echo box system on coupling method, pulse length, unloaded cavity $Q$, and transmitter-receiver properties has been developed. The power taken by a direct-coupled echo box has been plotted as a function of the ratio of loaded to unloaded $Q$.

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# A Network Theorem and Its Application*

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**Summary—The proof of a theorem useful in analyzing the response of certain types of circuits to modulated signals and examples of the use of the theorem are presented.**

## INTRODUCTION

IT IS KNOWN that the response of a high-frequency circuit to an amplitude-modulated sinusoidal input signal may sometimes be obtained by analyzing an "equivalent" low-frequency circuit whose input is the envelope only. Those who have used such analyses in the literature[1,2] probably are aware of the proof of the theorem presented here. However, the writer has not seen such a proof, and it appears to be in order to show in detail the basis for these excellent works. Therefore a theorem will be stated, its proof given, and examples of its use will be shown. The Laplace transformation and its inverse will be used as the most convenient means of presenting the material.

*Theorem:* Let $Z_H(s)$ be the transfer function of a high-frequency circuit normally operated at an angular frequency in the vicinity of $w_0$. Let $\zeta = s - jw_0$, where $s$ is the usual "complex frequency" and, of course, $\zeta$ is also complex.

Then, if

$$Z_H(s) = Z_L(s - jw_0) = Z_L(\zeta) \tag{1}$$

is the transfer function of some low-frequency circuit, the response of this low-frequency circuit to a driving force $m_i(t)$ is identical to the envelope of the response of the high-frequency circuit to a sinusoid of angular frequency $w_0$ whose envelope is $m_i(t)$.

To enlarge on these statements, let the input to $Z_H(s)$ be

$$i(t) = m_i(t)e^{jw_0t},$$

where it is understood as in steady-state alternating-current circuit analysis that the real, or imaginary, part of $i(t)$ is the physical driving force, and let its response to $i(t)$ be

$$e(t) = m_0(t)e^{jw_0t}. \tag{2}$$

These are shown in Fig. 1(a). The theorem postulates that if (1) holds, the output of $Z_L(\zeta)$ will be $m_0(t)$ for an input $m_i(t)$, as shown in Fig. 1(b).

*Proof:* Suppose that the Laplace transform of $m_i(t)$ is $M_i(s)$. The transform of $i(t)$ is then

$$I(s) = \int_0^\infty e^{-st}m_i(t)e^{jw_0t}dt$$

$$= \int_0^\infty e^{-(s-jw_0)t}m_i(t)dt = M_i(s - jw_0). \tag{3}$$

The Laplace transform of the response of the high-frequency circuit to $i(t)$ is given by the product of the transform of its input and its transfer function:

$$E(s) = I(s)Z_H(s), \tag{4}$$

while the time-response may be found by means of the inversion integral, provided it exists. That is,

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}E(s)ds, \tag{5}$$

where $\gamma > Re(s)$ for all poles of the integrand.

Using (3) and (4) in (5) we have

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}M_i(s - jw_0)Z_H(s)ds$$

or in view of (1)

$$e(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{ts}M_i(s - jw_0)Z_L(s - jw_0)ds.$$

Changing the variable of integration to $s - jw_0$,

$$e(t) = \frac{e^{jw_0t}}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t(s-jw_0)}M_i(s-jw_0)Z_L(s-jw_0)d(s-jw_0).$$

Now using the fact that $\zeta = s - jw_0$, we have finally

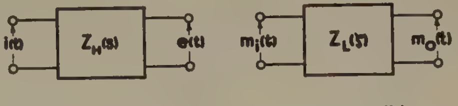
$$e(t) = e^{jw_0t}\left[\frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t\zeta}M_i(\zeta)Z_L(\zeta)d\zeta\right].$$

On comparing this last expression to (2), it is obvious that the coefficient of $e^{jw_0t}$ is the envelope of the response, namely

$$m_0(t) = \frac{1}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty} e^{t\zeta}M_i(\zeta)Z_L(\zeta)d\zeta. \tag{6}$$

We see that the right-hand side of (6) is also the response of the circuit characterized by $Z_L(\zeta)$ to $m_i(t)$, which proves the theorem.

*Example* 1. *A Simple Tuned Circuit.*[3] The simplest example of the use of the theorem is the analysis of an



(a) (b)

Fig. 1—(a) Showing the input and output of the high-frequency circuit. (b) Showing the input and output of the low-frequency circuit.

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* Decimal classification: R143. Original manuscript received by the Institute, May 18, 1950; revised manuscript received, March 5, 1951.

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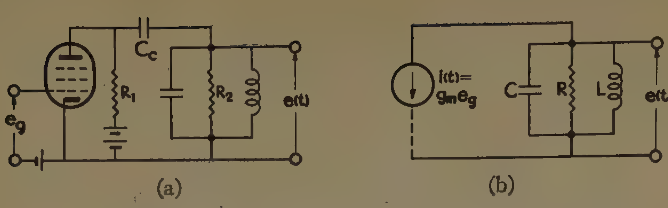
[1] G. E. Valley, Jr., and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., New York, N. Y., chap. 7; 1948.

[2] Eugene F. Grant, "Time response of an amplifier of $N$ identical stages," PROC. I.R.E., vol. 36, pp. 870–871; July, 1948.

[3] Grant's paper finds the response for $n$ stages of amplification, all stages tuned to the same frequency.



(a) (b)

Fig. 2—(a) The tube-amplifier circuit. (b) The equivalent circuit.

amplifier whose plate load impedance is a simple tuned circuit. (See Fig. 2.) The operational impedance of the tuned circuit is

$$Z_H(s) = \frac{R}{1 + Q_0\left(\frac{s}{w_0} + \frac{w_0}{s}\right)},$$

where

$$\frac{1}{R} = \frac{1}{r_p} + \frac{1}{R_1} + \frac{1}{R_2},$$

and if $|\zeta| \ll w_0$, this becomes

$$Z_H(s) \approx \frac{R\pi B}{\zeta + \pi B},$$

where

$$Q_0 = w_0 RC$$

$$w_0 = \frac{1}{\sqrt{LC}}$$

and

$$B = \frac{1}{R(2C)},$$

where $B$ is the bandwidth between the 3-db points of $|Z_H(s)|$. Thus

$$Z_H(s) \approx \frac{1/(2C)}{\zeta + \frac{1}{R(2C)}} = Z_L(\zeta). \qquad (7)$$

We see that (7) gives the operational impedance of the circuit shown in Fig. 3. It is such a simple matter to analyze this equivalent circuit for any reasonable $m_i(t)$, that no details will be put down here. It should be noted, however, that it is the approximation to $Z_H(s)$ in (7) which is equal to $Z_L(\zeta)$. We expect the results of an analysis of the equivalent circuit to be strictly accurate, then, only when $|\zeta|$ is indeed much less than $w_0$, but this is not a fault of the theorem.

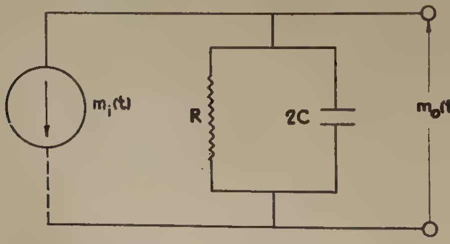


Fig. 3—The low-frequency equivalent of a simple tuned circuit.

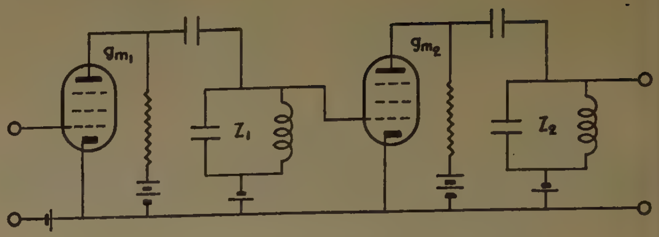


Fig. 4—A two-stage staggered-tuned amplifier.

Having obtained the equivalent low-frequency circuit, the reader might be tempted to include the components $R_1$ and $C_c$ (see Fig. 2(a)) ahead of those in Fig. 3, since these components may not be neglected at the frequencies contained in $m_i(t)$. However, when it is remembered that the frequencies dealt with in the original problem are close to $w_0$, at which the reactance of $C_c$ may be truly negligible, this pitfall is avoided.

*Example 2. Two Staggered-Tuned Stages of Amplification.* Consider the circuit of Fig. 4. In view of the previous section, the high-frequency equivalent circuit of each stage of amplification is like that of Fig. 2(b). If we denote the plate-load impedance of the first stage by $Z_1(s)$ and that of the second by $Z_2(s)$, then it is apparent that the gain of the two-stage amplifier is

$$Z_H(s) = g_{m_1} g_{m_2} Z_1(s) Z_2(s). \qquad (8)$$

Supposing that the two tuned circuits have equal bandwidths and are tuned respectively to angular frequencies $w_{0_1}$ and $w_{0_2}$, they have impedances which are approximately

$$Z_1(s) \approx \frac{\pi B R_1}{(s - jw_{0_1}) + \pi B}$$

and

$$Z_2(s) \approx \frac{\pi B R_2}{(s - jw_{0_2}) + \pi B},$$

where it has been assumed that $|s - jw_{0_1}| \ll w_{0_1}$ and $|s - jw_{0_2}| \ll w_{0_2}$. The magnitudes of these impedances (for $Re(s) = 0$, $Im(s) = w$) plotted versus $w$ will look something like the curves shown in Fig. 5.

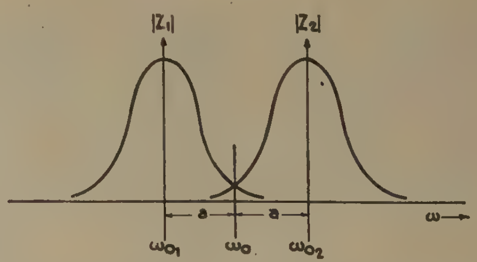


Fig. 5—The response curves of the staggered-tuned circuits.

Letting

$$w_{0_1} = w_0 - a$$

and

$$w_{0_2} = w_0 + a$$

the impedances may be written

$$Z_1(s) = \frac{\pi BR_1}{s - j(w_0 - a) + \pi B} = \frac{\pi BR_1}{\zeta + \pi B + ja} \quad (9)$$

$$Z_2(s) = \frac{\pi BR_2}{s - j(w_0 + a) + \pi B} = \frac{\pi BR_2}{\zeta + \pi B - ja} \cdot (10)$$

On inserting (9) and (10) in (8), we have

$$Z_H(s) = \frac{\pi^2B^2g_{m_1}R_1g_{m_2}R_2}{(\zeta + \pi B + ja)(\zeta + \pi B - ja)} = Z_L(\zeta). \quad (11)$$

Equation 11 suggests the low-frequency equivalent circuit of Fig. 6, where

$$R = \frac{\pi^2B^2g_{m_1}R_1g_{m_2}R_2}{\pi^2B^2 + a^2}$$
$$L = \frac{\pi Bg_{m_1}R_1g_{m_2}R_2}{2(\pi^2B^2 + a^2)^2} \quad (12)$$
$$C = \frac{2(\pi^2B^2 + a^2)}{\pi Bg_{m_1}R_1g_{m_2}R_2}.$$

For experimentally analyzing the response of odd-shaped modulations, this circuit might be of some value; however, to find the response of the amplifier to a unit step-function envelope is an easy matter. The inversion integral may be used to find $m_0(t)$ from (11):

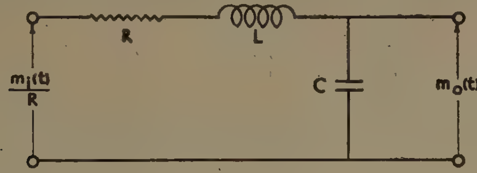


Fig. 6—The low-frequency equivalent circuit of a staggered-tuned amplifier.

$$m_0(t) = \frac{g_{m_1}R_1g_{m_2}R_2}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty}\frac{\pi^2B^2e^{t\zeta}d\zeta}{(\zeta + B + ja)(\zeta + B - ja)}.$$

We see from (11) that for $\zeta = 0$ (which means the circuit is excited by a sinusoid of angular frequency $w_0$) the gain is $g_{m_1}R_1g_{m_2}R_2/1+(a/\pi B).^2$ Letting $m_0'(t)$ be the normalized response, then,

$$m_0'(t') = \frac{m_0(t)[1 + \alpha^2]}{g_{m_1}R_1g_{m_2}R_2}$$
$$= \frac{1 + \alpha^2}{2\pi j}\int_{\gamma-j\infty}^{\gamma+j\infty}\frac{e^{t'\xi}d\xi}{(\xi + 1 + j\alpha)(\xi + 1 - j\alpha)}, \quad (13)$$

where, by a variable change, $\xi = \zeta/\pi B$ and $\alpha = a/\pi B$; that is, $\xi$ is the complex angular frequency in units of half-bandwidths of either tuned circuit, and $\alpha$ is the difference between the resonant angular frequency of either tuned circuit and the band center, also in half-bandwidths. Also $t' = \pi Bt$ is in units of half-bandwidth-seconds, or in radians.

We easily find $m_0'(t)$ by summing the residues of the poles of the integrand of (13). These poles occur at $\xi_1 = 0$, $\xi_2 = -(1+j\alpha)$, and $\xi_3 = -(1-j\alpha)$.

$$R_{\xi_1} = \frac{1}{1 + \alpha^2}$$
$$R_{\xi_2} = \frac{(1 - j\alpha)e^{-(1+j\alpha)t}}{2j\alpha(1 + \alpha^2)}$$
$$R_{\xi_3} = -\frac{(1 + j\alpha)e^{-(1-j\alpha)t'}}{2j\alpha(1 + \alpha^2)}$$

or

$$m_0'(t) = (1 + \alpha^2)\sum R$$
$$= 1 - e^{-t'}\left(\frac{1}{\alpha}\sin\alpha t' + \cos\alpha t'\right), \quad (14)$$

showing that the envelope of response of the staggered-tuned amplifier to a step-function modulated input first rises above and then oscillates about its final value. The damping of the oscillations depends on the bandwidth of either tuned circuit; this certainly is expected since the bandwidth itself is determined by the circuit time constant. The angular frequency of the oscillations of the envelope is equal to the separation of the resonant frequency of either circuit from the band center.

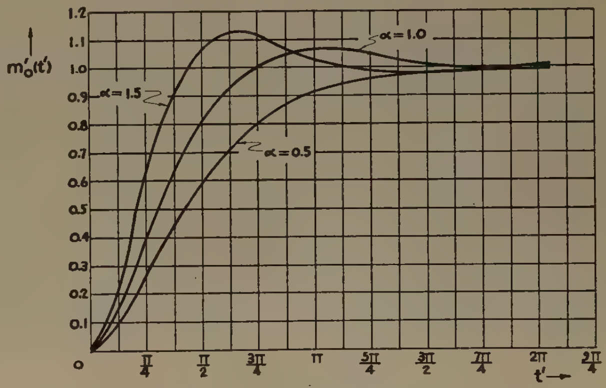


Fig. 7—The response of a staggered-tuned amplifier to a step-modulated signal.

Curves of $m_0'(t)$ versus $t'$ for several values of $\alpha$ are shown in Fig. 7. The points where $m_0'(t)$ reaches its maxima and minima are $t' = n\pi/\alpha$, the maxima being

$$m_0'(n\pi/\alpha) = 1 + e^{-(n\pi/\alpha)},\ n \text{ odd}$$

and the minima

$$m_0'(n\pi/\alpha) = 1 - e^{-(n\pi/\alpha)},\ n \text{ even}.$$

These are shown in Fig. 7.

The magnitude of the first "overshoot" is then $e^{-\pi/\alpha}$; obviously this is zero only when $\alpha = 0$, and in general it is smaller as $\alpha$ becomes small. For the maximally flat response curve,[4] $\alpha = 1$, the overshoot is $e^{-\pi} = 0.044$. If a small value of the overshoot is to be a criterion for choosing $\alpha$, then this should, of course, be as small as is consistent with a usefully large bandwidth, the maximally flat response curve providing a good comparison in this respect.

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[4] The response curve, that is, $|Z_H(s)|$ plotted versus $Im(s) = w$ has only one maximum, and no minimum for $\alpha = 1$.

From the curves we also see that the time for the response first to reach its final level is larger as $\alpha$ becomes smaller. This also follows from the fact that for $m_0'(t)$ to be equal to unity,

$$t' = \frac{1}{\alpha} \tan^{-1} (-\alpha).$$

If quick response is to be the criterion for choosing $\alpha$, then $\alpha$ should be as large as is allowed by overshoot considerations and steady-state response-curve flatness, $\alpha$ greater than unity resulting in the familiar double-humped response curve.

### CONCLUSIONS

The theorem provides a powerful means of analyzing the response of certain types of circuits to amplitude modulated signals. The step function has been used as a representative type of modulation, but the response to sinusoids, ramp functions, and the like, are easily investigated using the same methods.

The writer is at present attempting to extend the theorem to include modulated signals which are periodic, but not necessarily sinusoidal, and some success in this direction has been achieved. It is hoped that more general methods of analyzing circuits using modulated signals will be the subject of another paper.

# Constant-Resistance Networks of the Linear Varying-Parameter Type*

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*Summary*—**The term "constant-resistance network" is generally used to describe a fixed network whose input impedance is a constant independent of frequency. The present paper generalizes the concept of constant-resistance networks by extending this concept to linear varying-parameter systems. It is shown that any self-dual variable network is a constant-resistance network and this result is used to obtain such constant-resistance structures as the lattice and the bridged-tee. Explicit expressions for the transfer functions of these structures are derived and an illustrative example involving a constant-resistance variant of Cowan's modulator is considered. In addition to self-dual structures, a non-self-dual type of constant-resistance network is described and the possibility of using this structure as a constant-resistance capacitive modulator is indicated.**

## INTRODUCTION

THE TERM "constant-resistance network" is generally used to describe a fixed network whose input impedance is a constant independent of frequency. Such networks play an important role in the synthesis of four-terminal networks and, in particular, are widely used in the design of equalizers.[1]

The present paper generalizes the concept of constant-resistance networks by extending this concept to linear varying-parameter networks. This class of networks constitutes the most general form of linear systems; it includes as special cases the various kinds of amplitude and phase modulators, as well as the ordinary fixed linear networks.

As a preliminary to the consideration of variable[2] constant-resistance networks it is necessary to become acquainted with certain definitions and relations which form the basis for the circuit analysis[3] of variable networks in the frequency domain. These are as follows:

### *A. System Function*

The fundamental concept on which the circuit analysis of variable networks is based, is that of the system function[4] $H(j\omega; t)$. Briefly, the system function of a variable network $N$ is defined as a function $H(j\omega; t)$ such that $H(j\omega; t)e^{j\omega t}$ represents the response of $N$ to $e^{j\omega t}$. Thus, let $u(t)$ denote the input to $N$ and let $v(t)$ be the response to $u(t)$. Then, the definition of $H(j\omega; t)$ reads

$$H(j\omega; t) = \left. \frac{v(t)}{u(t)} \right|_{u(t)=e^{j\omega t}}. \qquad (1)$$

The salient property of the system function is that it relates the output and input of $N$ in much the same manner as the conventional system function $H(j\omega)$ relates the output and input of a fixed network. For example, the response of $N$ (at rest) to an input $u(t)$ ($u(t) = 0$ for $t<0$) may be expressed in terms of $H(j\omega; t)$ as

$$v(t) = L^{-1}\{H(s; t)U(s)\}, \qquad (2)$$

where $s=\sigma+j\omega$ is the complex frequency, $U(s)$ is the Laplace transform of $u(t)$, and $L^{-1}$ represents the operation of inverse Laplace transformation (treating $t$ in $H(s; t)$ as if it were a constant parameter). Similarly, in the operational form, one may write

$$v(t) = H(p; t)u(t), \qquad (3)$$

where $H(p; t)$ should be treated as a usual Heaviside operator, i.e., $t$ in $H(p; t)$ should be regarded as a con-

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† Columbia University, New York, N. Y.

[1] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., chap. 12; 1945.

[2] The term "variable," as used here and elsewhere in this paper, is synonymous with "linear varying-parameter."

[3] L. A. Zadeh, "Circuit analysis of linear varying-parameter networks," *Jour. Appl. Phys.*, vol. 21, pp. 1171–1177; November, 1950.

[4] L. A. Zadeh, "Frequency analysis of variable networks," PROC. I.R.E., vol. 38, pp. 291–299; March, 1950.

stant parameter. When written as $H(p; t)$, the system function will be referred to as a time-dependent Heaviside operator, or simply, as an operator.

In dealing with electrical networks it is convenient to refer to a system function by a distinctive name, such as impedance, admittance, gain, etc., which places in evidence the physical significance of $u(t)$ and $v(t)$. For instance, the *impedance* of a two-terminal network $N$ is in effect the system function relating the voltage across $N$, $v(t)$, to the current flowing through $N$, $i(t)$. Thus, denoting the impedance of $N$ by the symbol $Z(j\omega; t)$, (1) becomes

$$Z(j\omega; t) = \left.\frac{v(t)}{i(t)}\right|_{i(t)=e^{j\omega t}}. \tag{4}$$

Similarly, the *admittance* of $N$ is defined by the relation

$$Y(j\omega; t) = \left.\frac{i(t)}{v(t)}\right|_{v(t)=e^{j\omega t}}. \tag{5}$$

As a simple illustration of these definitions consider a variable capacitance $C = C(t)$, in which case the relation connecting $v(t)$ and $i(t)$ is

$$i = \frac{d(Cv)}{dt} \cdot \tag{6}$$

Applying (4) and (5), and using $s$ in place of $j\omega$, the expressions for the impedance and admittance of a variable capacitance are found to be

$$Z(s; t) = \frac{1}{C(t)s} \tag{7}$$

and

$$Y(s; t) = C(t)s + \dot{C}(t), \tag{8}$$

where the dot represents differentiation with respect to time.

### B. Product Relation

The product relation to be given below expresses essentially the rule of composition of successive operations on a function of time. Thus, let $v(t)$ be the result of successive operations by two operators, say $H_1(p; t)$ and $H_2(p; t)$, on $u(t)$. In other words,

$$v(t) = H_2(p; t)\{H_1(p; t)u(t)\}. \tag{9}$$

The two successive operations involved in (9) may be replaced by a single operation with an operator $H_3(p; t)$,

$$v(t) = H_3(p; t)u(t), \tag{10}$$

where $H_3(p; t)$ is called the *product* of $H_2(p; t)$ and $H_1(p; t)$. It can readily be shown that $H_3(p; t)$ is related to $H_2(p; t)$ and $H_1(p; t)$ through an operational expression

$$H_3(s; t) = H_2(p + s; t)H_1(s; t), \tag{11}$$

in which $H_2(p+s; t)$ is the operator and $H_1(s; t)$ plays the role of a function of time involving $s$ as a parameter. Equation (11) may conveniently be expressed in a symbolic form

$$H_3 = H_2 * H_1, \tag{12}$$

where the symbol $*$ has all the usual properties of an algebraic product except for commutativity. Thus, one can write

$$H_a * (H_b + H_c) \equiv H_a * H_b + H_a * H_c \tag{13}$$

and

$$H_a * (H_b * H_c) \equiv (H_a * H_b) * H_c, \tag{14}$$

where $H_a$, $H_b$, and $H_c$ represent arbitrary operators. However, in general, $H_a * H_b$ is not the same as $H_b * H_a$.

### C. Inverse of a System Function

The inverse of an operator $H(p; t)$, or simply $H$, is defined as an operator $H^{-1}$ such that the product of $H$ and $H^{-1}$ is equal to unity. Thus,

$$H * H^{-1} = H^{-1} * H = 1, \tag{15}$$

where the product symbol $*$ should be interpreted in the sense of (12). It should be noted that in general $H^{-1}$ *is not* the algebraic reciprocal of $H$. The determination of the inverse of a given time-dependent operator is generally a difficult problem. There is no need, however, to be concerned with this problem in the present paper.

Combining the definitions of the impedance and admittance of a variable network ((4) and (5)) with that of the inverse of a system function, it is readily seen that $Y(p; t)$ is the inverse of $Z(p; t)$, and vice versa. From this it follows that the impedance and admittance of a variable network are related to each other by the operational relation

$$Z(p + s; t)Y(s; t) = Y(p + s; t)Z(s; t) = 1 \tag{16}$$

or more simply

$$Z * Y = Y * Z = 1. \tag{17}$$

This relation represents a generalization of the familiar algebraic relation

$$Z(s)Y(s) = 1, \tag{18}$$

which holds in the case of fixed networks.

There are three important identities involving system functions which will be needed later in this paper. These are:

$$(H_a * H_b * H_c)^{-1} \equiv H_c^{-1} * H_b^{-1} * H_a^{-1} \tag{19}$$

$$\begin{aligned} H_a^{-1} + H_b^{-1} &\equiv H_a^{-1} * (H_a + H_b) * H_b^{-1} \\ &\equiv H_b^{-1} * (H_a + H_b) * H_a^{-1} \end{aligned} \tag{20}$$

$$\begin{aligned} (H_a^{-1} + H_b^{-1})^{-1} &\equiv H_b * (H_a + H_b)^{-1} * H_a \\ &\equiv H_a * (H_a + H_b)^{-1} * H_b. \end{aligned} \tag{21}$$

The properties of system functions as outlined above will now be used to extend the concept of constant-resistance networks to linear varying-parameter systems.

## CONSTANT-RESISTANCE NETWORKS

A variable network $N$ will be said to be a constant-resistance network if its input impedance $Z(s; t)$ is independent of both $s$ and $t$—that is, if $Z(s; t)$ is equal to a constant $k$. It is evident that if $Z(s; t)=k$ then $Y(s; t)=1/k$, and hence a constant-resistance network can be so characterized on the basis of either its impedance or its admittance. In what follows, the constant $k$ will be assumed to be equal to unity. This simplifies the discussion without entailing any loss in generality.

An important class of constant-resistance networks can be obtained by making use of the properties of dual variable networks. Duality in the case of variable networks involves the same relationships as in the case of fixed networks. The only difference is that in the case of variable networks the relations between the dual elements assume a more general form, which is as follows:

*The dual of an inductance $L=f(t)$ is a capacitance $C=f(t)$, and vice versa.*

*The dual of a resistance $R=h(t)$ is a conductance $G=h(t)$, and vice versa.*

It will be recalled that if $N$ and $N'$ are two fixed networks which are the duals of each other, then the input impedance of $N$ is equal to the input admittance of $N'$, and vice versa. The same relation holds in the case of variable networks. Thus, let $Z(s; t)$ and $Y(s; t)$ denote the input impedance and admittance of $N$, and let those of $N'$ be denoted by the corresponding primed symbols. Then

$$Z(s; t) = Y'(s; t) \tag{22}$$

and

$$Y(s; t) = Z'(s; t). \tag{23}$$

From the above relations it follows at once that any self-dual network—that is, any network which is identical with its dual—is a constant-resistance network. For, if $N$ is identical with $N'$, then (22) and (23) give

$$Z(s; t) = Y(s; t). \tag{24}$$

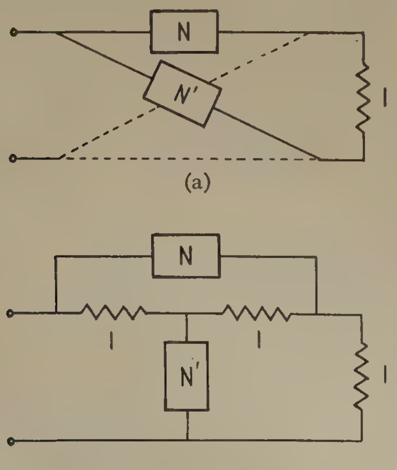

Fig. 1—Self-dual variable networks. (a) Constant-resistance lattice; (b) constant-resistance bridged-tee.

But, $Z(p; t)$ is the inverse of $Y(p; t)$, and therefore (24) implies that $Z(p; t)$ is equal to its inverse, which is impossible unless $Z(p; t)$ is equal to unity. Thus, it may be concluded that: *Any self-dual variable network is a constant-resistance network.*

This result may be used as a means for obtaining many forms of constant-resistance structures. For example, consider the lattice and the bridged-tee shown in Figs. 1(a) and 1(b), in which $N$ and $N'$ are dual variable networks. It can readily be verified that both of these structures are self-dual. Hence, we may conclude that:

*The lattice and the bridged-tee are constant-resistance networks whenever the branches $N$ and $N'$ consist of dual variable networks.*

In view of the importance of the constant-resistance lattice and the constant-resistance bridged-tee in the design of conventional equalizers, it is of interest to examine in somewhat greater detail the characteristics of these structures for the case where $N$ and $N'$ are variable networks.

Let $G(s; t)$, or simply $G$, denote the gain of these structures and let $Z(s; t)$ and $Z'(s; t)$, or simply $Z$ and $Z'$, represent the input impedances of $N$ and $N'$. Then, by using (14), (19), (20), and (21), and noting that $Z'$ is equal to $Z^{-1}$, it is readily found that the gain of the lattice is given by

$$G = (1 - Z) * (1 + Z)^{-1}, \tag{25}$$

where $(1+Z)^{-1}$ is the inverse of $(1+Z)$. This relation is a generalization of the well-known expression

$$G = \frac{1 - Z}{1 + Z}, \tag{26}$$

which represents the gain of a *fixed* constant-resistance lattice.

From (26), $Z$ may be expressed in terms of $G$ as follows:

$$Z = (1 - G) * (1 + G)^{-1}, \tag{27}$$

which is a generalization of

$$Z = \frac{1 - G}{1 + G}. \tag{28}$$

For a constant-resistance bridged-tee the corresponding relations are

$$G = (1 + Z)^{-1} \tag{29}$$

and

$$Z = G^{-1} - 1. \tag{30}$$

As a simple illustration of the use of the above relations, consider the case where it is desired that $G$ be a function of time of the form shown in Fig. 2(a). Applying (27) and noting that $Z'$ (the impedance of $N'$) is equal to $Z^{-1}$, it is easily found that $Z$ and $Z'$ should be variable resistances varying with time in the manner shown in Figs. 2(b) and 2(c). In this way one arrives at the constant-resistance variant of Cowan's modulator which was suggested by Tucker[5] on the basis of a different line of reasoning. It will be recognized, of course, that (27) and (30) make it possible to realize any desired form of modulating characteristic by using either a constant-resistance lattice or a constant-resistance bridged-tee.

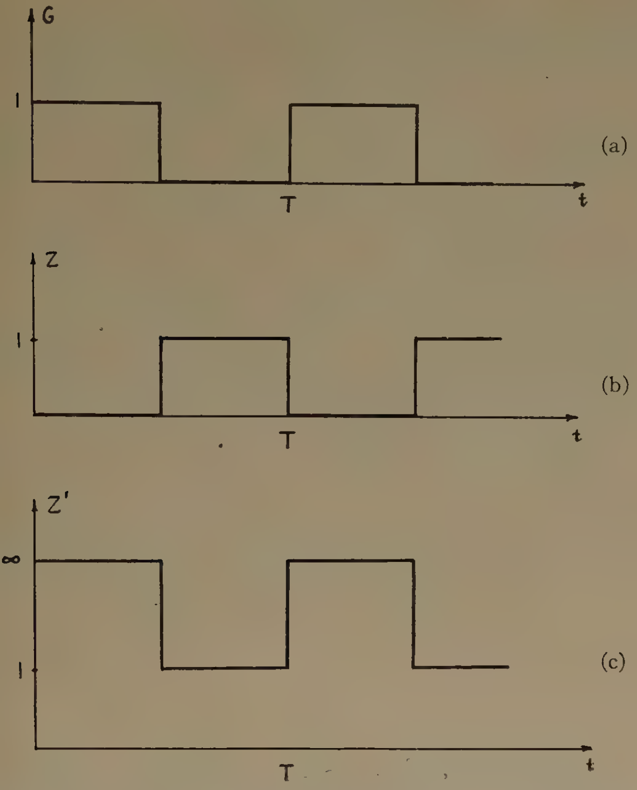

Fig. 2—(a) The desired variation of gain versus time; (b) variation of the impedance of series branches versus time; (c) variation of the impedance of crossed branches versus time.

It is evident that there are many forms of constant-resistance networks which are not of the self-dual type. The simplest of these is shown in Fig. 3. Here, as before, $N'$ is the dual of $N$. It will be noted that this structure was obtained simply by taking a well-known fixed constant-resistance network and replacing $N$ and $N'$ by variable networks. The same procedure can be used for obtaining variable constant-resistance networks of a more complicated form.

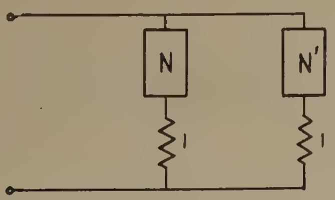

Fig. 3—One of the simplest non-self-dual constant-resistance networks.

To prove that the network shown in Fig. 3 is indeed a constant-resistance network, it is sufficient to show that its input admittance is equal to unity. Using the same notation as before, the input admittance in question may be written as

$$Y_{\text{in}} = (1 + Z)^{-1} + (1 + Z^{-1})^{-1}. \qquad (31)$$

By using (21), (31) becomes

$$Y_{\text{in}} = (1 + Z)^{-1} + Z * (1 + Z)^{-1} \qquad (32)$$

or

$$Y_{\text{in}} = (1 + Z)^{-1} * (1 + Z), \qquad (33)$$

and since the second factor in this expression is the inverse of the first factor, therefore $Y_{\text{in}}$ is equal to unity, which is the desired result.

For the particular case where $N$ is a capacitance $C = f(t)$, the network shown in Fig. 3 reduces to that shown in Fig. 4, in which the inductance $L'$ is the dual of $C$ and hence is equal to $f(t)$. The branch of this network comprising $C$ and $R(R = 1)$ may be regarded as a capacitive modulator of the type frequently used in practice. Thus, the network as a whole constitutes a constant-resistance capacitive modulator.

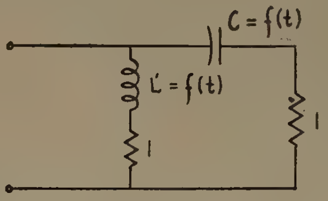

Fig. 4—A constant-resistance capacitive modulator.

Only a few basic types of constant-resistance networks have been considered in the preceding discussion. However, the outlined principles and methods are quite general and can be used both in the analysis and synthesis of various constant-resistance structures. In designing such structures it is useful to remember that the dual of any constant-resistance network is also a constant-resistance network, and that a complex constant-resistance structure may be obtained through the interconnection of a number of simpler constant-resistance networks.

## CONCLUSION

In conclusion, it may be stated that there is no essential difference between constant-resistance networks of the fixed and variable type. The potentialities of variable constant-resistance networks remain largely unexplored. They might prove useful in the design of constant-resistance modulators; in the simulation of time-dependent operators by means of electrical networks; in the synthesis of variable networks; and, as electromechanical analogs, in the design of constant-torque mechanical drives.

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[5] D. G. Tucker, "Rectifier resistance laws," *Wireless Eng.*, vol. 25, pp. 117–128; April, 1948. British Provisional Patent, 35540; 1946.

# Distortion in Linear Passive Networks*

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***Summary***—**It is well known that distortion is suffered by a modulated wave in passing through a practical transmission network.**

**The increasing use of frequency modulation in television relays and other uses where high fidelity is important has made the problem of FM distortion of current interest. One of the earliest analytical approaches to the problem was by Carson and Fry[1] and later applied by Jaffe[2] to single- and double-tuned circuits. The Carson and Fry method depends on a series solution to obtain numerical results.**

**The theory to be developed is based on a series solution of the superposition integral. A recent approach along this line is by Bradley.[3] He obtained a rapidly convergent series that is useful when the network transfer function is in a form which permits separation of attenuation and phase of the network and can be differentiated conveniently.**

**It is the purpose of this paper to provide a method of calculating distortion, when the transfer function is known in pole and zero form, that is rapidly convergent under most cases considered in practice.**

## THEORY

GIVEN A general input function $f_1(t)$, applied to a linear passive network $Y(\omega)$, the Fourier integral expression for the output is

$$f_2(t) = \int_{-\infty}^{\infty} f_1(\tau)\left[\frac{1}{2\pi}\int_{-\infty}^{\infty} Y(\omega)e^{j\omega(t-\tau)}d\omega\right]d\tau. \qquad (1)$$

The integral in the brackets will be recognized as the network response to a unit spike impressed at time $t$. This is usually called the characteristic transient[3] and denoted by the symbol $\text{Ŧ}\,(t-\tau)$.

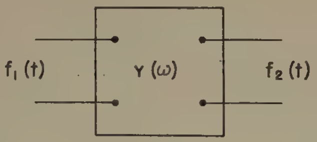

Fig. 1—General linear passive network.

Noting that $\text{Ŧ}(t-\tau)=0$ for $\tau>t$ in any physically realizable case, the output function $f_2(t)$ can be written as

$$f_2(t) = \int_{-\infty}^{t} \text{Ŧ}(t-\tau)f_1(\tau)d\tau. \qquad (2)$$

Now assume an input waveform

$$f_1(t) = Me^{j\omega_0 t}, \qquad (3)$$

where $M$ is the modulating function and is a function of time. The output waveform may be considered to be

$$f_2(t) = e^{\alpha+j(\omega_0 t+\beta)}, \qquad (4)$$

where $\alpha$ and $\beta$ are functions of $M$.

The general expression for the characteristic transient of a system of poles and zeros can be written as follows:[4]

$$\text{Ŧ}_t = \sum_n \sum_r \left[A_{nr}t^{(r-1)}e^{-(P_{nr}-j\omega_0)t} + A_{nr}{}^*t^{(r-1)}e^{-(P_{nr}{}^*+j\omega_0)t}\right], \qquad (5)$$

where

$A_{nr}{}^*$ is the conjugate of $A_{nr}$, $P_{nr}{}^*$ is the conjugate of $P_{nr}$, and

$$P_{nr} = \alpha_{nr} - j(\beta_{nr} - \omega_0) = \alpha_{nr} - j\sigma_{nr}.$$

From pole and zero theory, the summation over $n$ denotes $n$ different poles and the summation over $r$ denotes the $n$th pole is raised to the power $r$.

Combining (2), (3), (4), and (5),

$$e^{\alpha+j\beta} = \sum_n \sum_r t^{(r-1)}A_{nr}e^{-P_{nr}t} \cdot \int_{-\infty}^{t} M(\tau)e^{P_{nr}\tau}\left(1-\frac{\tau}{t}\right)^{(r-1)}d\tau + A_{nr}{}^*e^{-(P_{nr}{}^*-j2\omega_0)t} \cdot \int_{-\infty}^{t} M(\tau)e^{(P_{nr}{}^*-j2\omega_0)\tau}\left(1-\frac{\tau}{t}\right)^{(r-1)}d\tau. \qquad (6)$$

If the percentage bandwidth is small the second term in (6) becomes very small compared to the first term, and can be neglected. If wide percentage bandwidth is used (i.e., >12 per cent), the second term should be included and the following method is equally applicable. Equation (6) can now be rewritten for the narrow-band case,

$$e^{\alpha+j\beta} = \sum_n \sum_r A_{nr}t^r \int_0^{\infty} M(y)e^{-P_{nr}ty}y^{(r-1)}dy,$$

$$y = 1 - \frac{\tau}{t}\cdot \qquad (7)$$

At this point it is necessary to know the form of the modulating function $M(y)$ in order to apply this equation

### 1. *Input FM Wave*

For an input FM wave, the modulating function takes the form

$$M(\tau) = e^{j\phi(\tau)}, \quad \phi(\tau) = \phi[t(1-y)] = \theta(y).$$

Expanding $e^{j\theta(y)}$ in a power series in $y$,

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[*] Decimal classification: R148.11×R143. Original manuscript received by the Institute, January 19, 1950; revised manuscript received, August 22, 1950.

[†] Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Pa.

[1] J. Carson and T. Fry, "Variable frequency electric circuit theory with application to the theory of frequency-modulation," *Bell Sys. Tech. Jour.*, vol. 16, p. 513; October, 1937.

[2] D. Jaffe, "Tuned circuit F-M distortion," PROC. I.R.E., vol. 33, pp. 318–333; May, 1945.

[3] W. Bradley, "A Theory of FM Distortion," Philco Research Report #114.

[4] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y., vol. I, pp. 159–164; 1945.

$$e^{j\theta(y)} = e^{j\theta(o)} \sum_{s=0}^{\infty} A_s y^s, \quad \text{where}$$

$$A_s = \frac{1}{s!}\left[\frac{d^s}{dy^s}\left(e^{j\theta(y)}\right)\right]_{y=0} e^{-j\theta(o)}. \tag{8}$$

Combining (7) and (8) and performing the integration,[5] then

$$e^{\alpha+j\beta} = (B_0 + B)e^{j\phi(t)}, \tag{9}$$

where,

$$B_0 = \sum_n \sum_r A_{nr} A_0 \frac{(r-1)!}{P_{nr}^r},$$

$$B = \sum_n \sum_r \sum_{s=1} A_{nr} A_s \frac{t^{-s}(r-1+s)!}{P_{nr}^{r+s}}.$$

Although this form is frequently used to demonstrate distortion, it is not very useful in practice for application to distortion measurements. Taking the logarithm of both sides and letting $\log B_0 = \alpha_0 + j\beta_0$,

$$\alpha + j\beta = \alpha_0 + j[\beta_0 + \phi(t)] + \log(1 + B/B_0). \tag{10}$$

$\alpha_0$ and $\beta_0$ are independent of $\phi$, and therefore represent the attenuation and phase shift of the carrier in the absence of modulation. As they are constants, these two factors do not cause distortion. The last term is a function of the derivatives of $\phi$ and an expansion of this term will yield the distortion. From Appendix A,

$$B_0 + B = \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{(r-1)!}{(1-R)^r}\left[1 + j\frac{\phi''}{2P_{nr}{}^2} \cdot \frac{r(r+1)}{(1-R)^2} - j\frac{\phi'''}{6P_{nr}{}^3} \cdot \frac{r(r+1)(r+2)}{(1-R)^3} - \frac{\phi''^2}{8P_{nr}{}^4} \cdot \frac{r(r+1)(r+2)(r+3)}{(1-R)^4} + \cdots\right]. \tag{11}$$

For deviation ratios greater than one, this series converges very rapidly, the first term alone usually giving sufficient accuracy in practice. Although only the first term will be used in the remainder of this paper, the second term will provide a criterion for the accuracy of this method. Then,

$$B_0 + B \cong \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{(r-1)!}{(1-R)^r} \tag{12a}$$

$$B' = \sum_n \sum_r \frac{A_{nr}}{P_{nr}{}^r} \frac{r!R'}{(1-R)^{r+1}}$$

$$B' = \frac{dB}{dt} \quad \text{and} \quad R' = \frac{dR}{dt}. \tag{12b}$$

Returning to (10), the output frequency modulation is,

$$\beta' = \phi' + \Omega_D = \phi' + \text{imag.}\left[\frac{B'}{B_0 + B}\right], \tag{13}$$

where $\Omega_D$ is the output frequency-modulation distortion. The output amplitude-modulation distortion is,

$$M_D = e^\alpha - 1 = |B_0 + B| - 1. \tag{14}$$

The output wave can then be written,

$$f_2(t) = (1 + M_D)e^{j(\phi + \int \Omega_D dt)}. \tag{15}$$

The form of the above expression is convenient for application. If the above waveform is impressed on an ideal linear AM detector, the output is

$$f_{\text{AM}} = 1 + M_D. \tag{16}$$

The output of an ideal FM detector is,

$$f_{\text{FM}} = \phi' + \Omega_D. \tag{17}$$

It is profitable at this point to examine a typical FM receiver using a discriminator type detector circuit as illustrated by Fig. 2 in block diagram.

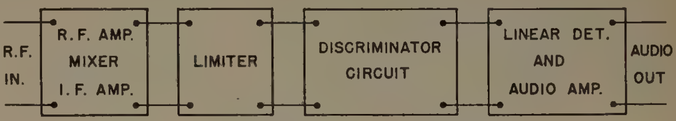

Fig. 2—Block diagram of typical FM receiver.

If an undistorted FM wave is applied to the receiver, this wave will suffer both AM and FM distortion in passing through the amplifier stages before the limiter. Assuming an ideal limiter, the AM distortion will be removed before the signal is applied to the discriminator. Thus, the FM distortion is of primary importance when considering the radio-frequency and intermediate-frequency amplifiers.

If a pure FM wave is applied to the discriminator, both FM and AM distortion are again produced by the tuned circuits. The output of the discriminator is applied to a linear detector that is assumed to be independent of frequency. Therefore, the AM distortion is of primary interest for this case. After detection, the fundamental frequency term in the AM distortion provides the useful output of a discriminator, and the higher harmonic terms provide the audio harmonic distortion.

In general, discriminator circuits are either one- or two-pole circuits (when the narrow-band approximation is used). The amplifier part of the receiver usually contains a larger number of poles. It will, therefore, be desirable to be able to calculate the amplitude distortion for one- or two-pole circuits and the FM distortion for any reasonable number of poles.

At this point it is convenient to consider special cases.

### *A. Synchronous Single-Tuned Amplifier.*

For this case all stages have equal damping and are tuned to the same frequency (thus, $n=1$ and $P_{1r}=P$). It is general practice to design the over-all rf and IF bandwidth to be equal to or greater than twice the maximum frequency deviation. For cases where this holds and the deviation ratio is sensibly greater than one, (12a) and (12b) can be used without appreciable error. If the percentage bandwidth is small, the term of major importance in the expression for the characteristic transient is the $r$th germ. For this case, using (12a), (12b), and (13), the general expression for the frequency distortion becomes,

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[5] H. B. Dwight, "Tables of Integrals," 861.2.

$$\Omega_{Dr} = \text{imag.}\left[\frac{rR'}{1 - R}\right] = \text{imag.}\left[\frac{-jr\phi''}{P+j\phi'}\right]. \qquad (18)$$

It should be noted that the above form of the expression for FM distortion does not require the amplifier to be tuned to the carrier frequency of the incoming signal. The only requirement is that the $r$ poles all be tuned to the same frequency and have the same damping.

Assume the modulating wave is a sinusoid, $\phi' = \Delta\omega \cos \omega_a t$. Rewriting (18) for this case, expanding into a Fourier series (see Appendix B), and with some trigonometric manipulation,

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \left(\frac{m_b}{m_T}\right)^m \sin m(\theta_1 - \theta_3) \sin m\omega_a t, \qquad (19)$$

where

$$m_b = \frac{\Delta\omega}{\sqrt{\alpha^2 + \sigma^2}}, \qquad \theta_1 = \tan^{-1}\frac{\alpha}{\sigma}$$

$$\theta_2 = \tfrac{1}{2}\tan^{-1}\frac{m_b^2 \sin 2\theta_1}{1 - m_b^2 \cos 2\theta_1}$$

$$\theta_3 = \tan^{-1}\left(\frac{\sin \theta_2 \sqrt[4]{1m_b^4 - 2m_b \cos 2\theta_1}}{1 + \cos\theta_2 \sqrt[4]{1m_b^4 - 2m_b \cos 2\theta_1}}\right)$$

$$m_T e^{j\theta_3} = 1 + \sqrt{1 - X^2}.$$

The above equation is in a convenient form for calculating purposes. It should be noted that considerable simplification occurs when $\sigma = 0$ (i.e., the circuit is tuned to the incoming carrier), and when $\sigma = \alpha$.

Several interesting general conclusions can be made from inspection of the above equation.

1. The distortion amplitude is linearly related to the modulation frequency.
2. For a given bandwidth per stage, the distortion increases linearly with the number of stages.
3. As there is no dc term, the incoming carrier is not shifted in frequency.
4. Only sine wave terms are present. The fundamental term represents "phase" distortion of the FM modulation, and the higher terms represent harmonic distortion.

Items 3 and 4 are surprising at first glance, as the carrier is not necessarily at the center frequency of the amplifier. The dc and cosine terms are, of course, present in the more rigorous general solution.

A special case of (19) will be considered:

1. Circuits tuned to incoming carrier, $\sigma = 0$. For this case

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \left(\frac{m_b}{1 + \sqrt{1 + m_b^2}}\right)^{2m-1} \cdot \sin(2m-1)\omega_a t, \qquad (20)$$

where $m_b = \Delta\omega/\alpha$.

As might be expected, only odd harmonics are produced.

If the over-all bandwidth of the amplifier is maintained constant, the value of $m_b$ increases with the number of stages. From synchronously tuned amplifier circuit theory,[6] it can be shown that

$$\omega_B = 2\alpha_1 = 2\alpha_0\sqrt{2^{1/r} - 1}, \qquad (21)$$

where $\alpha_1$ is the $\alpha$ for a one-stage single-tuned amplifier of bandwidth $\omega_B$. $\alpha_0$ is the $\alpha$ for an $r$-stage single-tuned amplifier of over-all bandwidth $\omega_B$.

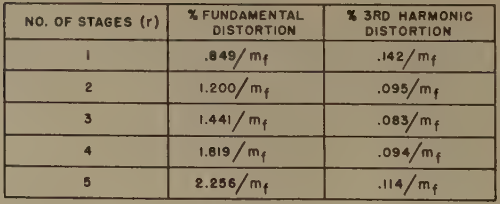
Combining (20) and (21),

$$\Omega_{Dr} = 2r\omega_a \sum_{m=1}^{\infty} (-1)^m \frac{\dfrac{2\Delta\omega}{\omega_B}\sqrt{2^{1/r} - 1}}{1 + \sqrt{\left(\dfrac{2\Delta\omega}{\omega_B}\right)^2 (2^{1/r} - 1) + 1}} \cdot \sin(2m-1)\omega_a t. \qquad (22)$$

Usually, in practice, the over-all bandwidth $\omega_B$ is equal to twice the maximum frequency of deviation. For this case, setting $\omega_B = 2\Delta\omega$, the percentage harmonic distortion becomes

$$\frac{\Omega_{Dr}}{\Delta\omega} = \frac{2r}{m_f} \sum_{m=1}^{\infty} (-1)^m \left(\frac{2^{1/2r} - 1}{2^{1/2r} + 1}\right)^{m-1/2} \sin(2m-1)\omega_a t, \qquad (23)$$

where $m_f = \Delta\omega/\omega_a$ is the frequency deviation ratio as usually defined. The absolute values of the amplitudes of the percentage harmonic distortion are plotted in Fig. 3 for several values of $r$. It should be noted that the fundamental distortion term can have a larger amplitude than the desired signal for $m_f$ near one. This term contributes both amplitude and phase distortion. The third harmonic distortion is very small for large values of $m_f$, and for values of $m_f$ near one, the third harmonic is usually outside the pass band of the audio amplifier. Thus, the fundamental distortion term is the most troublesome for this case.



| NO. OF STAGES (r) | % FUNDAMENTAL DISTORTION | % 3RD HARMONIC DISTORTION |
|---|---|---|
| 1 | .849/m_f | .142/m_f |
| 2 | 1.200/m_f | .095/m_f |
| 3 | 1.441/m_f | .083/m_f |
| 4 | 1.819/m_f | .094/m_f |
| 5 | 2.256/m_f | .114/m_f |

Fig. 3—Absolute values of amplitudes of the percentage harmonic distortion.

Using (20) or (22), the percentage harmonic distortion at any harmonic can be easily computed for a multiple stage synchronous single-tuned amplifier.

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[6] S. N. Van Voorhis, "Microwave Receivers," McGraw-Hill Book Co., New York, N. Y., pp. 160–162; 1948.

The amplitude modulation in the output can be found from (12a) and (14), and using the value of $A$ given in Appendix C,

$$M_D + 1 = \left(\frac{g_m}{g}\right)^r \cdot \frac{1}{\left[\left(1 + \frac{1}{m_\sigma^2}\right)\left(1 + \frac{m_b^2}{2}\right)\right]^{r/2}} \cdot \frac{1}{[1 + C \cos \omega_a t + D \cos 2\omega_a t]^{r/2}}, \quad (24)$$

where

$$C = \frac{4 m_b \sigma^2}{(2 + m_b^2)\Delta\omega}, \qquad D = \frac{m_b^2}{2 + m_b^2}.$$

The writer was not able to obtain a Fourier series expansion for (24) with coefficients in closed form. Fortunately, most applications only require $r=1$, and for this case the coefficients can be obtained in the form of rapidly converging power series. This is illustrated in the next example.

### B. Stagger-Tuned Discriminator.

The transfer function between the input current $i(t)$ and either output voltage, $e_A$ or $e_B$, in Fig. 3, can be represented by a single pole transfer function, $n=1$ and $r=1$.[7] When used as a stagger-tuned discriminator circuit, one pole is tuned slightly higher, $+\sigma$, than the carrier and the other tuned slightly lower, $-\sigma$. If the secondary $Q$'s are equal, the detected voltages (see (16)) at terminals $A$ and $B$ are

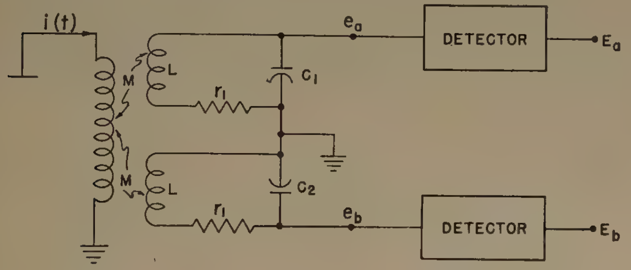

Fig. 4—Stagger-tuned discriminator.

$$E_A = M_D(+\sigma) + 1 \quad (25a)$$

$$E_B = M_D(-\sigma) + 1. \quad (25b)$$

The detected discriminator outputs are then subtracted in the usual manner to give the audio output $E_0$.

$$E_0 = E_A - E_B = M_D(+\sigma) - M_D(-\sigma). \quad (26)$$

From (24) for $r=1$,

$$M_D(+\sigma) + 1 = \left(\frac{g_m}{g}\right)^r \cdot \frac{1}{\sqrt{\left(1 + \frac{1}{m_\sigma^2}\right)\left(1 + \frac{m_b^2}{2}\right)}} \sum_{n=0}^{\infty} a_n \cos n\omega_a t, \quad (27)$$

where

$$a_0 = 1 + \frac{3}{16}(C^2 + D^2) - \frac{15}{64}C^2 D + \cdots$$

$$a_1 = \frac{1}{2}C + CD - \frac{15}{64}C^3 - \frac{15}{32}CD^2 + \cdots$$

$$a_2 = \frac{1}{2}D + \frac{3}{16}(C^2 + D^2) - \frac{15}{32}C^2 D + \frac{3}{4}D^3$$

$$a_3 = CD - \frac{5}{64}C^3 - \frac{15}{64}C^2 D.$$

It is clear that changing the sign of $\sigma$ changes the sign of $C$. Therefore, combining (25a), (25b), and (26),

$$E_0 = \left(\frac{g_m}{g}\right)^r \frac{1}{\sqrt{\left(1 + \frac{1}{m_\sigma^2}\right)\left(1 + \frac{m_b^2}{2}\right)}} \cdot \sum_{n=1}^{\infty} b_{(2n-1)} \cos (2n - 1)\omega_a t, \quad (28)$$

where, $b_1 = 2a_1$, $b_3 = 2a_3$, and so forth.

Two interesting facts are apparent:

1. All even harmonics are cancelled leaving only odd harmonics.
2. Amplitude distortion is present in the fundamental.

The amplitude of the fundamental would be proportional to $m_b$ if there were no distortion. In other words, the amplitude of the fundamental should be proportional to the frequency deviation. An examination of the fundamental Fourier coefficient $b_1$, shows that the term $C$ is the undistorted output, while the remaining terms give amplitude distortion. To minimize amplitude distortion, it is desirable to make $C$ much larger than $D$, but smaller than one. This is done most conveniently by choosing the value of $\sigma$ properly (i.e., $\sigma \geqq \Delta\omega$).

### 2. Input Amplitude-Modulated Wave

For an amplitude-modulated input, the modulating function takes the form,

$$M(\tau) = 1 + \cos \omega_a \tau, \quad M(y) = 1 + \cos \omega_a t(1 - y). \quad (29)$$

Inserting this expression for $M(y)$ into equation (7), and performing the integration[8]

$$\alpha + j\beta = \log \left[H_0 + H \cos (\omega_a t - \phi)\right] \quad (30)$$

where,

$$H_0 = \sum_n \sum_r \frac{A_{nr}}{P_{nr}^{\,r}}(r - 1)!$$

$$H_s = \sum_n \sum_r A_{nr}(r - 1)!\left(\frac{1}{P_{nr}^2 + \omega_a^2}\right)^{r/2} \cdot \sin \left(r \tan^{-1} \frac{\omega_a}{P_{nr}}\right)$$

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[7] L. Arguimbau, "Vacuum Tube Circuits," John Wiley and Sons, Inc., New York, N. Y., pp. 487–494; 1948.

[8] Bierens De Haan, "Nouvelles Tables D'Integrales Definis," G. E. Stechert, New York, N. Y., Tables 363, nos. 1 and 2, p. 507; 1939.

$$H_c = \sum_n \sum_r A_{nr}(r-1)!\left(\frac{1}{P_{nr}^2+\omega_a^2}\right)^{r/2} \cdot \cos\left(r \tan^{-1}\frac{\omega_a}{P_{nr}}\right)$$

$$H = \sqrt{H_c^2 + H_s^2}, \qquad \phi = \tan^{-1}\frac{H_s}{H_c} = \theta_\alpha + j\theta_\beta.$$

The output frequency modulation becomes

$$\beta' = \text{imag.}\left[\frac{-\omega_a R \sin(\omega_a t - \phi)}{1 + R\cos(\omega_a t - \phi)}\right], \text{ where } R = \frac{H}{H_0}. \qquad (31)$$

This is similar in form to (20), using the Fourier series expansion

$$\beta' = 2\omega_a \sum_{m=1}^{\infty} (-1)^m \left(\frac{R}{1+\sqrt{1-R^2}}\right)^m \cdot [\sin m(\theta_R - \theta_s) \sinh m\theta_\beta \cos m(\omega_a t - \theta_\alpha) - \cos m(\theta_R - \theta_s) \cosh m\theta_\beta \sin m(\omega_a t - \theta_\alpha)], \qquad (32)$$

where

$$m_r\epsilon^{j\theta_r} = R$$
$$m_s\epsilon^{j\theta_r} = 1 + \sqrt{1-R^2}.$$

It should be noted that if all the $\sigma_{nr}=0$, then $\beta'=0$. Thus $\beta'$ exists only when some poles are not tuned to the carrier frequency. This means that, in general, when an amplifier is not tuned to the carrier of the input AM wave, FM appears in the output wave.

The amplitude modulation for the general case is

$$M_A = \epsilon^\alpha = \left| H_0 + H\cos(\omega_a t - \phi) \right|. \qquad (33)$$

If $\sigma_{nr}=0$, $H$ and $H_0$ are real, and there is no harmonic amplitude distortion; only phase distortion is present.

It is convenient at this point to consider a special case:

1. Synchronous Single-Tuned Amplifier: If the percentage bandwidth is small, the term of major importance is the $r$th term. Assuming further that the amplifier is tuned to the incoming carrier, it is clear from (32) that $\beta'=0$. The amplitude modulation becomes,

$$M_A = \frac{A}{\alpha^r}(r-1)!\left[1 + \left(\frac{\alpha^2}{\alpha^2+\omega_a^2}\right)^{r/2} \cdot \cos\left(\omega_a t - r\tan^{-1}\frac{\omega_a}{\alpha}\right)\right]. \qquad (34)$$

Using this equation it is not difficult to calculate the output amplitude modulation for any value of $r$.

## CONCLUSION

The theory presented in this paper illustrates a method by which a wide variety of distortion problems can be analyzed. The multiple-stage synchronous single-tuned amplifier has been analyzed and the method can be extended to synchronous double-tuned amplifiers with a little added labor. Stagger-tuned amplifier problems can be handled without much difficulty, if the number of different pole positions is only two or three and symmetry exists around the midfrequency. Further application of the method to nonsinusoidal periodic modulating functions can be made by evaluation of (7) by methods similar to steady-state operational analysis.

In all examples worked out in this paper, the resultant distortion is found in terms of quantities that can be easily checked experimentally.

## ACKNOWLEDGMENT

The writer wishes to acknowledge the assistance of B. Haimowitz, Research Associate at the Moore School of Electrical Engineering, in obtaining several Fourier series expansions.

## APPENDIX A

### *Expansion of $B_0+B$*

In order to simplify notation, $p$ will be used to denote differentiation with respect to $y$ and $(p_n x)^m$ will signify the $n$th derivative of $x$ with respect to $y$ raised to the $m$th power. It will be understood that when $n$ or $m$ is negative, the expression is equal to zero. Then, from (8),

$$s!A_s\epsilon^{j\theta(o)} = [p^s\epsilon^{j\theta(y)}]_{y=0}$$
$$= \left[(pj\theta)^s + \frac{s(s-1)}{2}(pj\theta)^{s-2}(p^2j\theta) + \frac{s(s-1)(s-2)}{1\cdot 2\cdot 3}(pj\theta)^{s-3}(p^3j\theta) + \frac{s(s-1)(s-2)(s-3)}{2\cdot 4}(pj\theta)^{s-4}(p^2j\theta)^2 \cdots\right]_{y=0} \epsilon^{j\theta(o)}. \qquad (35)$$

Denoting differentiation with respect to $t$ by primes and noting that $p\theta(y) = -t\phi'(t)$, (35) can be combined with (8) and (9) to give

$$B_0 + B = \sum_n \sum_r \sum_{s=0} \frac{A_{nr}}{P_{nr}^r}\frac{(r-1+s)!}{s!} \cdot \left[R^s + j\frac{s(s-1)}{2}R^{s-2}\frac{\phi''}{P_{nr}^2} - j\frac{s(s-1)(s-2)}{1\cdot 2\cdot 3}R^{s-3}\frac{\phi'''}{P_{nr}^3} - \frac{s(s-1)(s-2)(s-3)}{2}R^{s-4}\frac{\phi''^2}{P_{nr}^4} + \cdots\right], \qquad (36)$$

where $R = -j(\phi'/P_{nr})$.

For $|R|<1$, the summations over $s$ can be performed and yield (11).

## APPENDIX B

The Fourier series expansion of

$$\frac{r\omega_a X \sin \omega_a t}{1 + X \cos \omega_a t}, \qquad X = j\frac{\Delta\omega}{p} \tag{37}$$

is obtained as follows:

$$\frac{1}{1 + X \cos \omega_a t} = \frac{a_0}{2} + \sum_{m=1}^{\infty} a_m \cos m\omega_a t, \; |X| < 1, \tag{38}$$

where

$$a_m = \frac{1}{\pi}\int_0^{2\pi} \frac{\cos mx dx}{1 + X \cos x} = \frac{1}{\pi}\int_0^{2\pi} \frac{\epsilon^{imx} dx}{1 + X \cos x}$$

let $Z = \epsilon^{ix}$, $dZ = iZdx$

$$a_m = \frac{1}{\pi X i}\oint \frac{Z^m}{Z^2 + \frac{2}{X}Z + 1}$$

$$= \frac{2(-1)^m}{\sqrt{1 - X^2}}\left(\frac{X}{1 + \sqrt{1 - X^2}}\right)^m. \tag{39}$$

The integral around the unit circle being evaluated at the two poles.

Combining (37), 39), and (20),

$$\Omega_{Dr} = \text{imag.}\left[2r\omega_a \sum_{m=1}^{\infty} (-1)^m \cdot \left(\frac{X}{1 + \sqrt{1 - X^2}}\right)^m \sin m\omega_a t\right]. \tag{40}$$

## APPENDIX C

*Characteristic Transient for an r-Stage Synchronously Tuned Amplifier.*

Using the narrow-band form for the Laplace transform of the admittance of a single-tuned circuit, i.e., $Y(p) = 2C(p - p_1)$, the characteristic transient for a synchronously tuned amplifier of identical stages is

$$\mathfrak{T}(t) = \mathcal{L}^{-1}\left[\left(\frac{g_m}{2C}\right)^r \cdot \frac{1}{(p - p_1)^r}\right], \tag{41}$$

where

$$p_1 = \alpha_1 - j\sigma_1.$$

$\mathcal{L}^{-1}$ indicates the inverse Laplace transform. Then,

$$\mathfrak{T}(t) = \left(\frac{g_m}{2C}\right)^r \frac{t^{r-1}}{(r - 1)!}\epsilon^{-p_1 t} \tag{42}$$

or,

$$A_r = \left(\frac{g_m}{2C}\right)^r \cdot \frac{1}{(r - 1)!}. \tag{43}$$

* * *

## CORRECTION

John Ruze, author of the paper, "Wide-Angle Metal-Plate Optics," which appeared on pages 53–59 of the January, 1950, issue of the PROCEEDINGS of the I.R.E., has advised the editorial department that Fig. 6 was in error.

Mr. Ruze therefore prepared the following illustration presented in the right-hand column, for the benefit of Proceedings readers who may have been confused. The editors are glad to co-operate in publishing this revised figure.

* * *

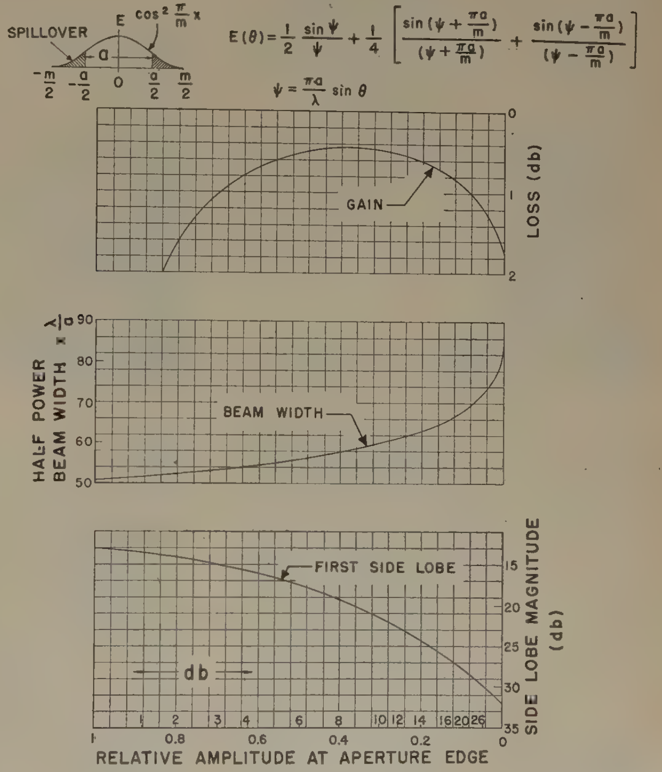

Fig. 6—Radiation characteristics of tapered rectangular aperture.

# Contributors to the Proceedings of the I.R.E.

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Richard C. Borden was born in Wilkes Barre, Pa., on April 9, 1919. After being graduated from Tulane University with the B.S. degree in physics in 1941, he was employed in the Division of Geomagnetism and Seismology of the United States Coast and Geodetic Survey. While on active duty in the Navy, he attended the Naval Officers' Radar Schools at Cruft Laboratory, Harvard University, and the Massachusetts Institute of Technology, and also served as radar officer aboard the aircraft carrier U.S.S. *Belleau Wood*, and as instructor at the Naval Officers' Pre-Radar School, Bowdoin College, Brunswick, Maine.

After separation from the Naval Service in 1946, Mr. Borden joined the engineering staff of the Civil Aeronautic Authority's Technical Development and Evaluation Center at Indianapolis, Ind. During this time he has been actively engaged in the development and evaluation of radar aids to air navigation and traffic control, particularly distance-measuring equipment. He is presently serving as Chief of the Radar Branch, Electronics Division.

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M. R. Boyd

Malcolm R. Boyd (S'50) was born on January 23, 1924, in Boulder, Colo. He received the B.S. degree in electrical engineering from the University of Colorado in 1947. In addition, he attended the University of Pittsburgh and Penn State College under the ASTP. From 1947 to 1949 Mr. Boyd was employed at the RCA Laboratories in Princeton, N. J., where he was engaged primarily in gaseous electronic studies. Since September, 1950, he has been a graduate student and research assistant in the electrical engineering department at Stanford University.

Mr. Boyd is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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For a photograph and biography of H. J. Reich, see page 1451 of the December, 1950, issue of the Proceedings of the I.R.E.

F. W. Bubb, Jr.

F. W. Bubb, Jr. (A'44) was born on July 12, 1921, at St. Louis, Mo. He received the B.S. degree in electrical engineering with honors in 1943, and the M.S. degree in applied mechanics in 1949, both from Washington University. From 1943 to to 1948 he was successively instructor at the Massachusetts Institute of Technology Radar School in Boston; chief product engineer at the Navy division of McQuay-Norris Manufacturing Company, St. Louis, Mo., a VT fuze assembly plant; senior design engineer at the Naval ordnance division of Eastman Kodak Company, Rochester, N. Y.; and research engineer engaged in work on automatic control devices at Washington University, St. Louis, Mo. During 1948, Mr. Bubb invented and has applied for patent on an electronic instrument for measuring and controlling fluid phases and extremely low humidities. In 1948, he joined the staff of St. Louis University as an assistant professor of engineering.

Mr. Bubb is a member of Tau Beta Pi, Sigma Xi, and Pi Mu Epsilon, and is a registered professional engineer in the fields of applied mechanics and electrical engineering.

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Georg Goubau

Georg Goubau (A'49) was born in Munich, Germany, on November 29, 1906. He received the Dipl. Phys. degree in 1930, and the Dr. Ing. degree in 1931, both from the Munich Technical University. From 1931 to 1939 he was employed in research and teaching in the physics department of the same university, under Professor Zenneck. During this time he was principally concerned with ionospheric investigations. He established the first German Ionospheric Research Station (Herzogstand/Kochel), and was in charge of the research work carried on at this station.

In 1939 Dr. Goubau was appointed professor and director of the department of applied physics at the Friedrich-Schiller University, in Jena, Germany. Before he arrived in this country, he was the senior author of the volumes on electronics of the FIAT Review of German Science, published by the Military Government for Germany. Dr. Goubau is now a consultant at the Signal Corps Engineering Laboratories, in Fort Monmouth, N. J.

William M. Hall

William M. Hall (A'33–M'44–SM'45) was born in Burlington, Vt., on July 10, 1906. He attended the University of Vermont and received the B.S. degree in electrical communications from the Massachusetts Institute of Technology in 1928.

From 1928 until 1930 he was with the Sound Picture Industry in New York, N. Y. Following this, he was on the staff of the electrical engineering department of MIT until the beginning of World War II. In 1934 he received the Sc.D. degree from that institution. During 1940 and 1941, Dr. Hall was on the staff of the MIT Radiation Laboratory.

Since 1941 he has been associated with the Raytheon Manufacturing Company, where he was a project engineer on various radar and countermeasures systems during the war and where he now holds the position of technical assistant to the manager of the radar department.

Dr. Hall is a member of Sigma Xi and a Fellow of the Acoustical Society of America.

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Norman L. Harvey

Norman L. Harvey (A'41–M'46–SM'48), was born in Ossian, Iowa in 1911. He was graduated from the Iowa State College in 1934 with the B.S. degree in electrical engineering. In 1941, Mr. Harvey joined the research department of Sylvania Electric Products, Inc., at Emporium, Pa., and was actively engaged in a number of developments for the armed services. In 1943 he was transferred to the Central Engineering department at Flushing, L. I., N. Y., as supervisor of the circuits section. Here his duties comprised responsibility for a number of developments including a special radar unit, a part of which was later used in the "moon radar" experiments, as well as in ultra-high-frequency communication equipment.

At present Mr. Harvey is director of engineering of the Colonial Radio and Television Division of Sylvania, in Buffalo, N. Y

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Meyer Leifer (A'46–M'48–SM'50) was born in New York, N. Y., in 1914. He received the B.S. and M.S. degrees from

Brooklyn College and Columbia University, respectively, the former in mathematics and the latter in physics. After teaching for several years in the New York City secondary schools, he joined the United States Navy and served as a radio-radar officer during the war. Mr. Leifer became a member of the physics laboratories staff of Sylvania Electric Products Inc. upon his release from active duty, and has been engaged principally in the application of the theory of communication to navigation and television systems.

Meyer Leifer

Mr. Leifer is now the leader of the theoretical group of the circuits section, and is teaching in the Graduate School of the Brooklyn Polytechnic Institute. He is a member of the American Physical Society.

◆

Donald Gordon Lindsay was born on May 2, 1909, in Australia, and was educated at North Sydney High School and Sydney University, where he completed a two-year Science course. He joined the Amalgamated Wireless (A/sia) Ltd. in 1933 as a design engineer, after previous experience with Stromberg-Carlson (A/sia) Ltd.

Donald G. Lindsay

Mr. Lindsay's association with the Institution of Radio Engineers, Australia, as a foundation member dates from 1932. He was elected a Fellow of this Society in 1949.

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Nean Lund (A'40) was born in Madison, Wis., on May 13, 1912. He was graduated in 1935 from the University of Wisconsin, where he received the B.Sc. degree in electrical engineering. After a year of postgraduate study at that University, and a brief period as a radio consultant, he joined the staff of the Bell Telephone Laboratories, Inc. He first worked on the development and testing of single-sideband radio receivers. From 1941 to 1943 he was concerned with circuit development of loran, radar, and other equipment used by the military.

Nean Lund

In recent years, Mr. Lund has worked on the measurement of interchannel modulation noise from multiplex single-sideband transmitters, and on the measurement of extraband radiation from multiplex single-sideband radio transmitters. He is a member of the American Institute of Electrical Engineers, Phi Eta Sigma, and Eta Kappa Nu.

◆

Louis Malter (A'37–SM'45) was born on April 28, 1907, in New York, N. Y. He received the B.S. degree from the College of the City of New York in 1926 and the M.A. and Ph.D. degrees in physics from Cornell University in 1931 and 1936, respectively. From 1926 to 1928 he taught physics at the College of the City of New York. He was with the Acoustic Research and Photophone Divisions of Radio Corporation of America from 1928 to 1930 and during the summer of 1931. From 1933 to 1942 Dr. Malter was associated with RCA Manufacturing Company at Camden, N. J., and from 1942 to 1943 with the RCA Laboratories Division at Princeton, N. J. During the period from 1943 to 1946, he was in charge of the Special Development Division of RCA Manufacturing Company at Lancaster, Pa. From May, 1946, to December, 1947, he was in charge of the Vacuum Tube Research Section of the Naval Research Laboratory in Washington, D. C. In December, 1947, he returned to the RCA Laboratories where he is now engaged in research on gaseous electronics.

L. Malter

Dr. Malter is a fellow of the American Physical Society, and a member of Sigma Xi.

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Nathan Marchand (A'39–M'44–SM'50) was born on June 20, 1916, in Shawinigan Falls, Canada. He received the B.S. degree from the College of the City of New York and the M.S. degree from Columbia University, both in electrical engineering.

Nathan Marchand

Mr. Marchand has been a lecturer in electrical engineering and has served as a consultant in the fields of cardiology and antenna and circuit design. He is at present the head of the circuits section of the physics laboratories, at Bayside, L. I., N. Y., as well as a consultant at the New York University-Bellevue Medical Center. He is the author of the books "Frequency Modulation," and "Ultrahigh Frequency Transmission and Radiation."

Mr. Marchand is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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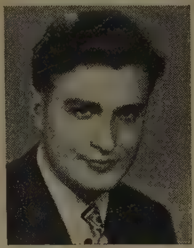
James A. Marsh (S'49–A'50) was born in Youngstown, Ohio, on January 7, 1922. He received the degree of A.B. from Ohio Wesleyan University in 1946, and both the M.Sc. and Ph.D. degrees from The Ohio State University in 1947 and 1949, respectively, where his field of specialization was electrical engineering.


J. A. Marsh

During the war, Dr. Marsh was an electronics officer in the Army Air Force, working in the field of radar. He taught for the year 1947–1948 on the electrical engineering department staff at The Ohio State University. From 1948 to 1950 he was associated with the Antenna Laboratory of The Ohio State University, where he did research on flush-mounted airborne antennas. In February, 1950, Dr. Marsh joined the staff of the Aerophysics Laboratory of North American Aviation, Inc. He is now directing research in the antenna and microwave section. Dr. Marsh is a member of Phi Beta Kappa, Eta Kappa Nu, Pi Mu Epsilon, and Phi Mu Alpha.

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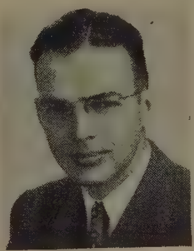
De Forrest Metcalf (M'48) was born in El Paso, Tex., on November 25, 1916. After receiving the B.S. degree from the University of Texas in 1937 he was employed by the Humble Oil and Refining Company. In 1942 he entered the employ of the Raytheon Manufacturing Company and participated in its microwave development program. He joined the staff of the Electrical Engineering Research Laboratory of The University of Texas in 1948, where he has been in charge of the wave propagation research conducted under contract with the National Bureau of Standards.


D. F. Metcalf

Mr. Metcalf is a member of Tau Beta Pi.

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Wilbur L. Pritchard (A'45–M'48) was born in New York, N. Y., on May 31, 1923. He received the B.E.E. degree from the College of the City of New York in 1943, and has since done graduate work at the Massachusetts Institute of Technology.

W. L. Pritchard

During the period from 1943 to 1946, Mr. Pritchard worked as an engineer for the Philco Radio and Television Corporation, where he was engaged in the development of airborne radar systems and later with the home radio and phonograph division. Since 1946 he has been associated with Raytheon Manufacturing Company, first as a development engineer in the microwave communications department and then as section head in charge of the radio-frequency and antenna section, his present position.

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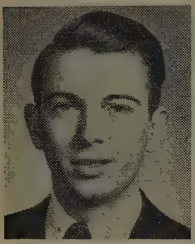

LOUIS E. RABURN

Louis E. Raburn (S'41–A'44–M'46–SM'50) was born in Manhattan, Kan. on June 4, 1919, and received the B.S. degree in electrical engineering from Kansas State College in 1941. From 1935 to 1941 he was employed as an engineer at radio station KSAC. From August, 1941, until March, 1942, he was employed as a carrier engineer with the long lines department of the American Telephone and Telegraph Company.

From March, 1942, until October, 1945, Mr. Raburn was a research associate at Radio Research Laboratory, Cambridge, Mass., and attended Harvard University graduate school.

Mr. Raburn has been with Electronics Research, Inc., since November, 1945. He has been engaged in the development of low-drag and zero-drag aircraft antennas, FM broadcast antennas, and antenna test equipment. In addition, as an antenna engineer, he has supplied consultant services to several aircraft companies on the application of zero-drag antennas. At present, Mr. Raburn is the senior engineer of the antenna division of Electronics Research, Inc.

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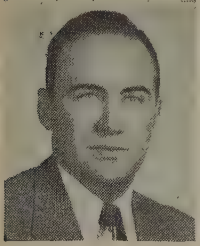

LEON RIEBMAN

Leon Riebman (S'43–A'45) was born on April 22, 1920, in Coatesville, Pa. He received the B.S. degree in electrical engineering in 1943, and the M.S. degree in 1947, both from the Moore School of Electrical Engineering, University of Pennsylvania.

After attending the United States Navy Midshipman School, he was assigned to duty in radar development at the Navy Research Laboratory in 1944. In March, 1946, he joined the research and development staff of Philco Corporation as a senior engineer. From October 1, 1948, until June, 1949, he attended the Moore School of Electrical Engineering, University of Pennsylvania, under an Atomic Energy Commission pre-doctoral fellowship. Since June, 1949, he has been associated with the Moore School as a Research Associate and part-time Instructor in electrical engineering.

Mr. Riebman is a member of Sigma Xi and Tau Beta Pi.

*Credit C. T. Alburtis*

P. M. SCHULTHEISS

Peter M. Schultheiss (S'50–A'51) was born in Munich, Germany, on October 18, 1924. He received his education in Germany, England, and the United States, where, in 1945, he received the B.E. degree from Yale University.

During the period from 1944 to 1946, Mr. Schultheiss served with the U. S. Army Signal Corps. Following this he returned to Yale for graduate work and received the M.E. degree in 1948. In that year he was appointed an instructor in electrical engineering at Yale, where, since 1950, he has been an assistant professor.

Mr. Schultheiss has specialized in the field of servomechanisms and has also been active in research on transistor circuits.

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GEORGE SINCLAIR

George Sinclair (A'37–SM'46) was born in Hamilton, Ont., Canada, on November 5, 1912. He received the B.Sc. degree in electrical engineering in 1933 and M.Sc. degree in 1935 from the University of Alberta, and the Ph.D. degree in 1946 from the Ohio State University. Dr. Sinclair was an instructor in electrical engineering at the University of Alberta for one year, and an engineer for the Northern Broadcasting Corporation for two years.

From 1941 to 1947 Dr. Sinclair was a research associate in the department of electrical engineering of the Ohio State University, supervising the research program of the Antenna Laboratory. He is now an associate professor of electrical engineering at the University of Toronto, and is also a consultant for the Antenna Research Laboratory Inc., Columbus, Ohio.

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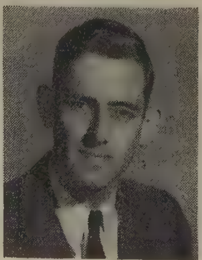

F. W. SCHOTT

Frederick William Schott (S'41–A'45–M'50) was born in Phoenix, Ariz., on October 2, 1919. In 1940, he received the degree of Bachelor of Arts (Honors) from San Diego State College. In 1941, he was awarded a university scholarship and began graduate work at Stanford University. The following year he continued under a Westinghouse Fellowship, and received the degree of electrical engineer in 1943. He resumed graduate study at Stanford University in 1947 and received the Ph.D. degree in 1948.

He has held the following positions: junior engineer, San Diego Gas and Electric Co., San Diego, Calif., 1943 to 1944; ensign, United States Navy, 1944 to 1946; instructor in engineering and physics, San Diego State College, 1946 to 1947; acting instructor, Stanford University, summer, 1947; research associate, Stanford University, 1947 to 1948; and assistant professor, University of California at Los Angeles, 1948 to date. During 1949 and 1950 he was on leave of absence with the U. S. Navy Electronics Laboratory.

Dr. Schott is a member of Tau Beta Pi, Sigma Xi, Sigma Pi Sigma, the American Association for the Advancement of Science, and the American Institute of Electrical Engineers, and is a registered electrical engineer in the State of California.

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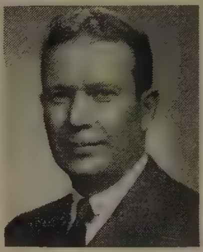

R. C. SPRAGUE

Robert C. Sprague was born on August 3, 1900, in New York, N. Y. He attended Hotchkiss, the United States Naval Academy and Post-Graduate School, and the Massachusetts Institute of Technology, where he also took postgraduate work during 1923–1924. He continued his naval career as a naval architect and later was a member of the staff which superintended the design and building of the aircraft carrier U.S.S. *Lexington.*

Mr. Sprague founded the Sprague Electric Company (originally the Sprague Specialties Co.), in 1926. In addition to being president and director of this company, he is a director of the Sprague Products Company of North Adams, Mass., and also of the Herlec Corporation, of Milwaukee, Wis. He actively participated in co-ordinating industry and government activities during World War II as a member of the WPS Advisory Committee on Electric Condensers, 1942–1945; chairman of the OPA Industry Advisory Committee for the Radio Parts Industry, 1944; and member of the Executive Committee of the Massachusetts Committee on Postwar Reconversion, 1942. He has been a Director of the Radio Manufacturers' Association since 1943, and has been chairman of the RMA Parts Division for two terms, 1944–1945 and 1945–1946. More recently he has been chairman of the RMA "Town Meetings" committee, which directed numerous activities in the interest of radio and television dealers and service technicians. He is now chairman of the board of Radio-Television Manufacturers' Association.

Mr. Sprague is a Fellow of the American Institute of Electrical Engineers, and also vice-president and member of the executive committee of the Associated Industries of Massachusetts.

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A. W. Straiton (M'47–SM'49) was born in Tarrant County, Tex., on August 27, 1907. He received the degrees of B.S. in electrical engineering in 1929, M.A. in 1931, and Ph.D. in 1939, all from the University of Texas.

A. W. Straiton

Dr. Straiton spent one year at Bell Telephone Laboratories, after which he taught at Texas College of Arts and Industries as assistant professor, associate professor, and professor of electrical engineering, successively. From 1941 to 1943, he was head of the Department of Engineering, institutional representative of Engineering, Science, and Management War Training, and director of the Army Pre-Radar Training Courses. Since 1943, he has been associate professor and later professor of electrical engineering at the University of Texas. He was associate director of the Electrical Engineering Research Laboratory from 1945 to 1947 and director from 1947 to the present.

Dr. Straiton is a Member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, the American Institute of Electrical Engineering, and the American Society for Engineering Education.

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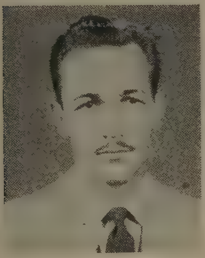
Charles Walter Tolbert (M'49) was born in Homer, La., on August 28, 1920. He received the B.S. degree in general engineering from Texas College of Arts and Industries in June, 1943. He taught in the Army Pre-Radar Training Program at Texas College of Arts and Industries and in the Wright Field Pre-Engineering Program at the University of Texas. For two years, he was engaged in geophysical exploration with National Geophysics Company.



C. W. Tolbert

In November, 1945, Mr. Tolbert joined the staff of the Electrical Engineering Research Laboratory of The University of Texas, where he is still employed as radio engineer in charge of constructing and testing of very-high-frequency and microradio equipment.

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Carl C. Trout (A'44–M'49) was born in Roanoke, Va., on March 31, 1916. He attended Bridgewater College in 1934 and 1935. Subsequently he was employed as an inspector in the War Department Office of the Chief Signal Officer in 1942, and attended the Signal Corps Radar School at Lexington, Ky. He was then assigned to the Air Materiel Command as radar installation engineer in 1943, later transferring to the Communications and Navigation Laboratory, in Dayton, Ohio, in 1944. Here he served as project engineer on microwave radar beacons and distance-measuring equipment.

Carl C. Trout

In 1947 Mr. Trout accepted his present position at the Civil Aeronautics Administration, Technical Development and Evaluation Center, at Indianapolis, Ind., where he has been engaged in the development and improvement of distance-measuring equipment.

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Henning E. von Gierke was born in Karlsruhe, Germany, on May 22, 1917. He studied electrical engineering and electronics at the Technical Universities of Karlsruhe and Munich. He received the degree of Diplom Ingenieur at Karlsruhe in 1943. From 1943 to 1947 he was research assistant at the Institute for Theoretical Electrical Engineering and Communications Technique at the Technical University of Karlsruhe. Here he worked on several projects in the fields of acoustics and applied physics. In 1944 he received the degree of Doctor Ingenieur, from the Technical University of Karlsruhe.

H. E. von Gierke

From 1946 to 1947 Dr. von Gierke served as lecturer in high-frequency communications technique at this institution. Since 1947 he has acted as a consultant in the Aero Medical Laboratory, Engineering Division, Air Materiel Command, Wright-Patterson Air Force Base, Dayton, Ohio. Fields of special interest to him are pressure, temperature, and vibration measurement problems; sound generation and radiation from air jets; propagation of sound in body tissues, and the physics of the ear.

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Wolf-Wito von Wittern was born in Dresden, Germany, on July 8, 1909. He received the degree of Diplom Ingenieur from the Technical University, Dresden, in 1935. From 1938 to 1943 he was a member of the Institute for Electrophysics at the Deutsche Versuchsanstalt für Luftfahrt, Berlin Adlershof. In 1943 he became head of a department of the Helmholtz-Institute, Nussdorf/Inn.

W.-W. von Wittern

Since 1947 Mr. von Wittern has worked as a consultant in acoustics and vibration at the Aero Medical Laboratory, Engineering Division, Air Materiel Command, Wright-Patterson Air Force Base, Dayton, Ohio. In his research he has chiefly been concerned with problems in the fields of acoustics, gas dynamics, and vibrations, as encountered in aviation and aerodynamics.

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Emory C. Williams was born on June 26, 1920, at St. Johns, Mich. He received advanced technical training while serving in the U. S. Navy at Texas A. & M. College and at Corpus Christi Naval Aviation Technical Training Center. Later, during his tour of active duty in the Navy, he served as an electronics technician in the Pacific theater. Following discharge from the Navy in 1946, he joined the staff of the CAA Technical Development and Evaluation Center at Indianapolis, Ind., where he was assigned as a project engineer in the Radar Branch. While with the TDEC, he was engaged in the development and evaluation of radar-type aids to air navigation and traffic control. During this period, he continued his engineering training at Purdue University Extension.

Emory C. Williams

Mr. Williams transferred to the CAA Aeronautical Center at Oklahoma City in September, 1950, and is presently serving as an instructor of distance-measuring equipment at the CAA training school associated with the Aeronautical Center.

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Lotfi A. Zadeh (S'45–A'47–M'50) was born on February 4, 1921, at Baku, Russia. He attended Alborz College of Teheran, which is an American missionary school and received the B.S. degree in electrical engineering from the University of Teheran in 1942. He worked for a year as a technical contractor with the U. S. Army Forces in Iran, and came to the United States in 1944. After a brief association with International Electronics Laboratories in New York, N. Y., he resumed his studies, receiving the M.S. degree from the Massachusetts Institute of Technology in 1946, and the Ph.D. degree from Columbia University in 1949. In 1946 he joined the staff of Columbia University, where he is now assistant professor of electrical engineering.

Lotfi A. Zadeh

Dr. Zadeh is an associate member of the American Institute of Electrical Engineers, and a member of the American Physical Society, the American Mathematical Society, and Sigma Xi.

# Correspondence

## Duality, a New Approach to Transistor Circuit Design*

A new approach to the problem of transistor circuit design has recently yielded very promising results. This approach is based on a recognition that there is a dual[1] relationship between the electrical properties of a transistor and those of a vacuum-tube triode. Because of this relationship, it is possible to start with a known vacuum-tube circuit and transform it into a completely different circuit suitable for use with transistors.

The duality between vacuum tube and transistor properties is illustrated by the static characteristics of Fig. 1. Note that the two sets of characteristics are quite similar, provided transistor potentials are compared with vacuum-tube currents and vice versa. In particular, the quantities to be compared are $i_c$ with $-v_p$, $v_c$ with $-i_p$, $i_e$ with $-v_g$, and $v_e$ with $-i_g$.

This comparison indicates that the transistor is approximately the dual of a vacuum tube. It is argued, therefore, that the dual of a vacuum-tube circuit should be suitable for use with transistors.

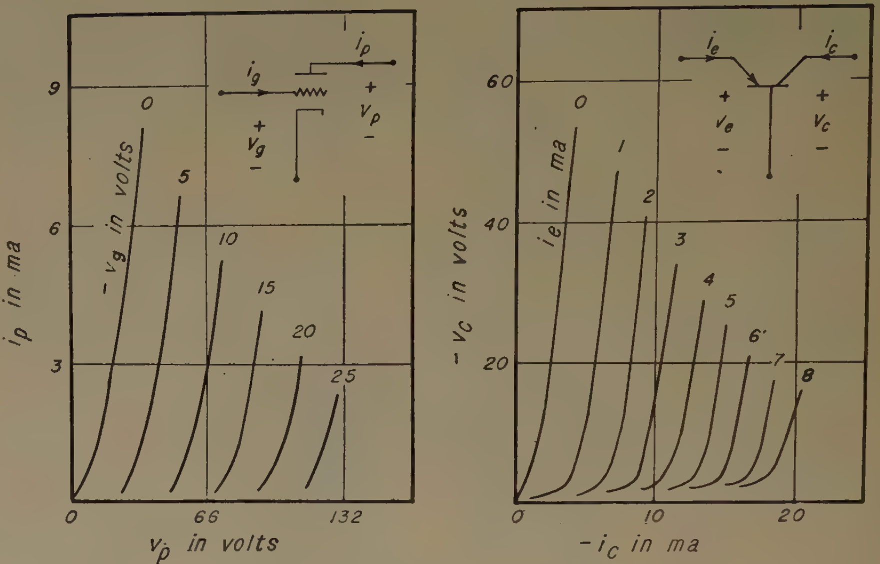

Fig. 1

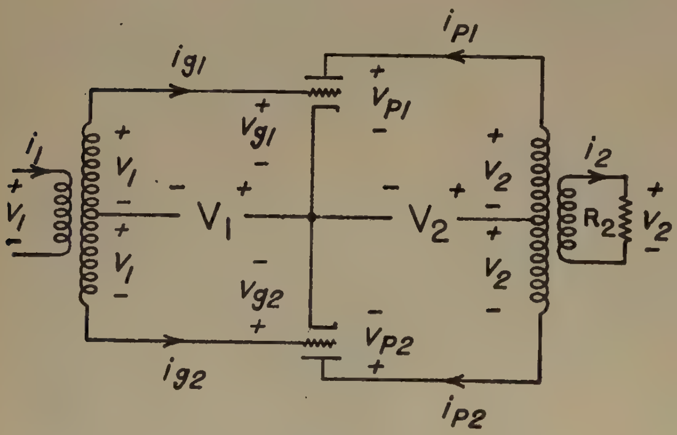

Fig. 2

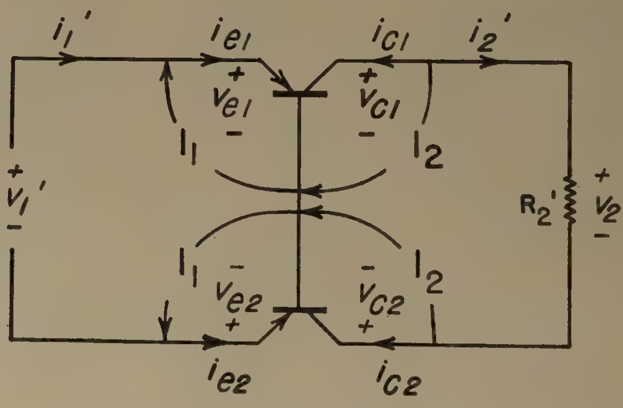

Fig. 3

As a particular example, consider the class-B amplifier of Fig. 2, the dual of which is shown in Fig. 3. The arrows labeled $I_1$ and $I_2$ indicate constant bias current supplies. For the dual of class-B operation the emitters are biased heavily forward, so that in the absence of a signal the collector potentials are reduced nearly to zero while the collector currents are high. In response to a positive half cycle of input current, the collector potential of the upper transistor remains essentially zero, while the collector potential of the lower transistor goes through a negative half cycle. During this half cycle the upper transistor is inactive, serving simply as a short-circuit element in series with the load. The corresponding situation in the vacuum-tube case is that the inactive tube is effectively an open-circuit element in shunt with the load.

With type *A* transistors in the circuit of Fig. 3, it has been possible to obtain 400-milliwatts output with reasonably low distortion and with collector circuit efficiency of the order of 60 per cent. If the same transistors are used in the conventional circuit of Fig. 2, the obtainable power output is very much lower—approximately 25 milliwatts.

Further investigation has shown that the methods of approach proposed here lead to many other useful transistor circuits.

R. L. Wallace, Jr.
Bell Telephone Labs., Inc.
Murray Hill, N. J.

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* Received by the Institute, January 25, 1951.

[1] E. A. Guillimin, "Communication Networks," John Wiley & Sons, Inc., New York, N. Y., vol. 2, p. 246; 1935.

# Institute News and Radio Notes

## TECHNICAL COMMITTEE NOTES

A Task Group on Pulse Definitions was set up under the **Standards Committee** recently, for the purpose of obtaining a set of definitions which would satisfy the needs of all IRE Technical Committees. This Task Group, headed by C. J. Hirsch of Hazeltine, has prepared the **Standards on Pulses: Definitions of Terms, 1951,** which is published in this issue of the PROCEEDINGS. J. G. Brainerd announced that arrangements have been made by the IRE Executive Committee to effect co-operation with the Institution of Radio Engineers in Australia on standardization work.

The Institute wishes to express its gratitude to Professor Brainerd for his leadership and guidance during the past two years as Chairman of the Standards Committee, and for his sincere interest and co-operation as IRE Representative on the ASA Standards Council. A. G. Jensen, the new Standards Committee Chairman, will replace Professor Brainerd on the ASA Council.

A meeting of the **Receivers Committee** was held on March 28 under the Chairmanship of R. F. Shea. This Committee has prepared two standards: namely, a proposed Standards on Radio Receivers: Open Field Methods of Measurement of Spurious Radiation, Frequency Modulation and Television Broadcast Receivers; and a proposed Standards on Receivers: Definitions of Terms. Both of these were submitted to the Standards Committee for approval on April 27.

The **Committee on Sound Recording and Reproducing** held a meeting on April 3, H. E. Roys, Chairman, at which the chairmen of the various subcommittees announced their tentative work schedules for the year. This Committee is concerned primarily with definitions of terms and measurement procedure in the fields of magnetic, mechanical, and optical recording.

A. Gardner Fox, Chairman of the **Antennas and Waveguides Committee,** presided at a meeting of that group on April 10. The Committee is drafting definitions of terms relating to waveguides and transmission lines.

R. E. Shelby and R. M. Bowie have been appointed Chairman and Vice-Chairman, respectively of the **Television Systems Committee.**

A meeting of the **Circuits Committee** was held on April 13, W. N. Tuttle, Chairman. Work on the preparation of definitions is progressing.

The **IRE Co-ordination Committee with International Technical Organizations (CCITO),** under the chairmanship of Axel G. Jensen, held a meeting on April 9 and proposed a question regarding a standard frequency-band designation to be put before the CCIR at its next convention in Geneva.

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## ATOMIC STANDARDS OF LENGTH NOW AVAILABLE

The availability to science and industry of an ultimate standard of length was announced today by the National Bureau of Standards and the Atomic Energy Commission. The standards consist of spectroscopic lamps containing a single pure isotope of mercury. These lamps enable any research organization which has the auxiliary optical equipment to have for the first time an ultimate primary standard of length in its own laboratories.

Distribution of the lamps will be handled by the National Bureau of Standards. They will be available to qualified government, industrial, and education laboratories, both in this country and abroad, engaged in precision length measurements and related research. All requests for information and applications should be addressed to the Co-ordinator of Atomic Energy Commission Projects at the National Bureau of Standards, Washington 25, D. C.

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## PROFESSIONAL GROUP NOTES

Two new Professional Groups were formed during the month of May: the **IRE Professional Group on Industrial Electronics,** and the **IRE Professional Group on Information Theory.** Two petitions for the formation of the **Industrial Electronics Group** were received; one from Eugene Mittelmann of Chicago, Ill., and one from Carl E. Smith of the United Broadcasting Co., Cleveland, Ohio. On May 22, this Group held its first conference in Cleveland, Ohio, sponsored jointly with the Cleveland Sections of the IRE and the AIEE. Eight technical papers were presented.

At least four Professional Groups have already decided to assess their members in order to raise money for the purposes of Group publications, or for conducting Group symposia. Up to $1,000 of each Group's funds will be matched by Institute funds, and the Groups will be well on their way toward financial independence.

The following newly organized Professional Groups have not yet had an opportunity to conduct a national membership drive:

**Airborne Electronics**
**Industrial Electronics**
**Information Theory**
**Radio Telemetry and Remote Control**

Membership application forms for these Groups are available upon request from IRE Headquarters.

The Editorial Department of the IRE has offered to serve the Professional Groups in the following ways: (1) it will endeavor to procure papers for Groups; (2) it will notify a Group when a paper of particular interest to it is received; (3) it will publish a boxed notice preceding a paper procured by a Group, stating that the paper originated with that Group, (4) upon submission of a Group paper for publication, it will give it expedited handling; (5) it will furnish suggested formats for Professional Group publications for both mimeographed or offset publications, and for the more formal Group TRANSACTIONS; and (6) it will place at the disposal of the Groups the experience of the Editorial Department in editing or making arrangements with printers.

A meeting of the Administrative Committee of the **IRE Professional Group on Airborne Electronics** was held in Dayton, Ohio, on April 5. Plans were made for the Group's participation in the Airborne Electronics Conference, which was held from May 23 to May 25 at the Biltmore Hotel in Dayton. The Group sponsored an evening symposium and a luncheon at the Conference, and held a business meeting of its members. A constitution and bylaws were drawn up.

Abstracts of papers presented at the URSI/IRE meeting held in Washington, D. C., on April 16–19 appear in this issue of the PROCEEDINGS. The Institute's participation was conducted by the **IRE Professional Group on Antennas and Propagation.** The Group hopes that these papers will later appear in complete form in a publication of the Group's TRANSACTIONS. A meeting of the Administrative Committee of the Group was held on April 18 in Washington, D. C., to discuss plans for participation in the IRE West Coast Convention which will take place in San Francisco, Calif., August 22–24. There will be three sessions sponsored jointly by the Group and the West Coast Convention.

The **IRE Professional Group on Audio** is currently making plans to publish two papers for its members.

A constitution and bylaws are being drawn up by the **Professional Group on Broadcast Transmission Systems.** The local Group in Boston held a meeting on March 8 in the studios of WCOP. Forty-five people attended, and Roger W. Hodgkins, Chairman, presided. A paper entitled, "Use and Interpretation of Noise and Distortion Meters," was presented by Charles Cady. A meeting has been held, also, in Troy, N. Y., to initiate a local Group in Northern New York State.

The **Professional Group on Circuit Theory** is drawing up a constitution and bylaws. Several IRE Sections have shown interest in the formation of local Circuit Theory Groups.

The **IRE Professional Group on Instrumentation** has a financial balance resulting from co-sponsorship of the High-Frequency Measurements Conference in January of this year. This will be matched by Institute funds in accordance with the new IRE policy. The following people have been elected as new members of the Group's Administrative Committee to serve a three-year term commencing July 1: John F. Byrne, Rudolf Feldt, W. D. Hershberger, and Hugh S. Knowles. Several members of the Group, under the leadership of W. M. Rust, Jr., Regional Director of IRE Region No. 6, will participate in the Sixth National Instrument Conference to be held in Houston, Texas, next fall, by organizing an IRE session on Instrumentation. The Group will organize a session on High-Frequency Measurements for the 1951 National Electronics Conference in Chicago, Ill., next October. The local Detroit Group, with 32 members under the Chairmanship of Seymour Sterling, has already scheduled several meetings, and its activities are well under way. Interest in the formation of additional local Groups exists in Chicago, Los Angeles, and in other areas.

The money earned by the **Professional Group on Nuclear Science** as a result of its Conference on Electronic Instrumentation in Nucleonics and Medicine will be retained by the Group and augmented by a similar amount from Institute funds in accordance with the new financial policy for Professional Groups. The Group is planning another conference, to be sponsored jointly with the AIEE, which will probably be held next fall. Urner Liddell will serve as Chairman of the Planning Committee for the conference. Ballots for the election of two new members of the Group's Administrative Committee have been mailed to the membership.

The **IRE Professional Group on Quality Control** has amended its constitution and bylaws, and has submitted them for approval. Election of six new members to the Group's Administrative Committee is being conducted. The Committee held a meeting on April 23 in Chicago, Ill., details of which will be reported in these Notes next month.

A membership drive is currently being conducted by the **Professional Group on Radio Telemetry and Remote Control.** The Group plans to publish two newsletters each year, to sponsor papers for the IRE West Coast Convention, and to hold a symposium on Telemetering on the West Coast in the spring of 1952.

Ballots have been mailed to members of the **Professional Group on Vehicular Communications** to elect five new members of the Administrative Committee. Plans are under way for the Group's technical meeting to be held in the fall. Reprints of five technical papers which have appeared in the Proceedings were furnished through the courtesy of Bell Telephone Laboratories, and mailed by IRE Headquarters to the Group's entire membership.

### Engineers Wanted!

Electrical and electronics engineers are urgently needed for work in research and development with the Armed Forces Security Agency. All positions are permanent insofar as this agency is concerned, and are located in the metropolitan area of Washington, D. C. Neither Civil Service examinations nor Civil Service status is required. Salaries range from GS-7, $3825.00, to GS-12, $6400.00 per annum, depending upon applicants' education and experience. For further information, write to the Director, Armed Forces Security Agency, Washington 25, D. C., Att'n: AFSA-153C.

### IRE Members, Eligible for Draft—Attention!

The Signal Corps has recently established the following procedures which will assist in the proper assignment of technically skilled IRE members upon entry into the Army:

The Institute of Radio Engineers will provide, upon request, to members who are about to be called into military service in enlisted grades, a Statement of Experience indicating the most applicable military specialties in the radio-electronic field for which a member's experience and background qualify him. This Statement of Experience should be presented by the member upon reporting for military service for reference in assignment to duty.

Interested members should write to George W. Bailey, Chairman, Liaison Committee, 1 East 79 St., New York 21, N. Y., giving full particulars as to education and experience in the radio-electronic field.

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## British IRE Plans Summer Convention

The British Institution of Radio Engineers is planning to hold its third convention this summer, during the Festival of Britain. The program will include the following sessions: "Electronic Instrumentation in Nucleonics," University College, London, July 3–4; "Valve Technology and Manufacture," University College, London, July 5–6; "Radio Communication and Broadcasting," University College, Southampton, July 24–25; "Radio Aids to Navigation," University College, Southampton, July 26–27; "Television Engineering," King's College, Cambridge, August 21–24; and "Audio-Frequency Engineering," The Richmond Hall, Earls Court, September 4–6.

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## Engineering Research Fellowships Offered

Research fellowships in electrical engineering will be awarded by the Institute of Industrial Research, University of Denver. The stipend is $3,300 for 21 months starting September 1, 1951. The recipients will work approximately 25 hours per week in the Institute of Industrial Research on industrial research projects leading to the fulfillment of the thesis requirement, while engaged in graduate study leading to the M.S. degree in electrical engineering. Interested applicants should send transcript of college record, personal data, and names of three persons acquainted with their work to the Director, Institute of Industrial Research, University of Denver, Denver 10, Col.

## Oak Ridge Institute to Give Additional Courses

Three additional basic courses in radioisotope technique of four-week duration, and a three-week autoradiography course will be offered this summer by the Special Training Division of the Oak Ridge Institute of Nuclear Studies.

The basic courses will be the 21st, 22nd, and 23rd to be given by the Institute since the program was initiated in June, 1948, and are scheduled to begin on June 11, July 9, and August 15. They are to combine lectures, demonstrations, and laboratory work. Sufficient space is available to permit individual laboratory work. Thirty-two participants will be accepted for each course.

The autoradiography course beginning on July 2 will be the first of its kind to be offered by the Institute. It is intended for personnel who will direct medical or biological research utilizing the autoradiographic process.

George A. Boyd, formerly of the University of Rochester, well-known for his work in autoradiography, will direct the course. Subjects to be covered include the theory of photographic process, reaction of ionizing particles with photographic emulsions and the interpretation of results, and techniques of making gross and microscopic autoradiograms. Twenty participants will be accepted for the course.

Lecturers and laboratory leaders of the course will include the following: Julian Webb and John Spence, Eastman Kodak Co.; Robert Dudley, MIT; Margaret Holt, New England Deaconess Hospital; C. P. Leblond and Rita Bogorach, McGill University; L. F Belanger, University of Ottawa S. R. Pelc, Hammersmith Hospital, London; Agnes Williams, University of New Mexico; and Harvey Blank, University of Pennsylvania.

Registration is $25 per course. Additional information and application forms are available from Ralph T. Overman, Chairman, Special Training Division, Oak Ridge Institute of Nuclear Studies, P.O. Box 117, Oak Ridge, Tenn.

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### Attention!

### Firms in New York Area

The New York Section of the IRE is interested in arranging field trips of technical interest to its members to places conveniently located to New York City. It is contemplated that such trips be scheduled on Saturdays, and that about 100 or more persons participate.

Organizations in the communications, electronic, or allied fields able to accommodate an IRE field trip should contact A. L. Stillwell, Chairman of Field Trip Committee, IRE New York Section, Bell Telephone Laboratories, Inc., Murray Hill, N. J.

## Over 600 Attend New England Radio Engineering Meeting

On April 21 the Copley Plaza Hotel in Boston, Mass., was the scene of a very successful New England Radio Engineering Meeting, sponsored by the North Atlantic Region of the IRE. The all-day program of papers and exhibits attracted over 600 engineers, scientists, and educators from the six-state area.

The morning technical session was under the chairmanship of Herbert L. Krauss, Chairman of the Connecticut Valley Section. The following papers were presented: "Radio Frequency Problems in the Design of a Linear Accelerator," by H. L. Schultz and W. G. Wadey, Yale University, New Haven, Conn.; "Considerations in the Design of a Line of Inexpensive Test Equipment," by D. B. Sinclair, General Radio Co., Cambridge, Mass.; and "Spectrum Utilization in Color Television," by R. B. Dome, General Electric Co., Syracuse, N. Y.

Willard H. Hansen, Chairman of the Boston Section, presided over the afternoon session at which three more papers were read: "Physiological Effects of Radiation," by W. A. Meissner, New England Deaconess and Baptist Hospitals, Boston, Mass.; "Civilian Defense Against Atomic Attack," by J. W. M. Bunker, Massachusetts Institute of Technology, Cambridge, Mass.; and "Instrumentation in the Field of Radioactivity," by W. A. Higinbotham, Brookhaven National Lab., Upton, L. I., N. Y.

Following the afternoon session, a Regional meeting was held for IRE members with Regional Director H. J. Reich presiding.

Sharing the spotlight with the sessions was the informative exhibit of recent radio-engineering and electronic developments. The products and services of 101 firms were shown in an outstanding representation of the radio-electronic industry.

Social activities included a luncheon at which IRE President I. S. Coggeshall gave an address of welcome.

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## Pacific Coast Regional Conference Slates Program

The Seventh IRE Regional Conference, to be held at the University of Washington, Seattle, Wash., on June 20, 21, and 22, has announced the following program:

On Wednesday morning, June 20, registration will take place in the Student Union Building, after which a paper on color television will be read.

In the afternoon the following subjects will be considered: transformer design, new coupling circuits, loudspeaker design, optimum servo design, and computers.

On Thursday morning, June 21, papers will be read on such topics as telemetering in 12 mev, a low-frequency generator, a microsecond pulser, and impedance measurement to 1,000 mc.

The afternoon session will be devoted to multiple operation of transmitters and receivers, linear rf amplifier circuits, spectrophotometers, facsimile, cutoff waveguides, and a miniature hf amplifier.

The topics to be dealt with on Friday morning, June 22, are as follows: If ionosphere measurement, ground-wave phase shift, aircraft antenna design, VOR cavity antenna, a slot aircraft antenna, and signal distribution from broadcast DA.

In the afternoon theater television equipment, a 30-inch tv receiver, uhf television systems, a 5-kw uhf tv amplifier, and a micro-wave relay for television discussed.

In addition to these technical sessions, field trips are being planned to the Puget Sound Navy Yard on a destroyer escort, to the Boeing Aircraft Co., to the University of Washington's new cryclotron, to a television station, and other places of interest.

### Calendar of COMING EVENTS

54th Annual Meeting of the American Society for Testing Materials, Chalfonte-Haddon Hall, Atlantic City, N. J., June 18–22

**IRE 7th Regional Conference, University of Washington, Seattle, Wash, June 20–22**

**Electron Devices Conference, University of New Hampshire, Durham, N. H., June 21–22**

1951 AIEE Summer General Meeting, Royal York Hotel, Toronto, Ont., Canada, June 25–29

1951 Annual IAS Summer Meeting, 7660 Beverly Blvd., Los Angeles, Calif., June 27–28

Institute of Navigation National Meeting, Hotel New Yorker, New York, N. Y., June 28–30

International Isotope Conference, Oxford, England, July 16–21

**1951 IRE West Coast Convention, San Francisco, Calif., Aug. 22–24**

**1951 National Electronics Conference, Edgewater Beach Hotel, Chicago, Ill., October 22–24**

**Radio Fall Meeting, King Edward Hotel, Toronto, Ont., Canada, October 29–31**

**1952 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6**

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## Scientists Wanted For Civil Service Positions

The United States Civil Service Commission has called attention to the continuing need in the Federal Service for chemists, metallurgists, physicists, mathematicians, and electronic scientists. Applications are being accepted for positions in and around Washington, D. C., paying from $3,825 to $10,000 a year.

No written test will be given to applicants for these positions. To qualify, they must meet a basic requirement of appropriate college education and/or experience. In addition, they must show that they have had professional experience in the field for which they apply. These and other requirements are explained in detail in the examination announcement.

Information and applications may be obtained at most first- and second-class post offices, from Civil Service regional offices, or from the United States Civil Service Commission, Washington 25, D. C. Applications are to be sent to the Executive Secretary, U. S. Civil Service Committee of Expert Examiners, National Bureau of Standards, Washington 25, D. C., where they will be accepted until further notice.

# Industrial Engineering Notes[1]

### Controls

Defense Production Administrator William H. Harrison has announced the **establishment of an Electronics Production Board** which will be responsible for over-all coordination of electronics production under the mobilization program. DPA officials stated that the board "will consist of a Chairman to be appointed by the Administrator, the Director of the Electronics Division of the National Production Authority, the General Manager of the Atomic Energy Commission, representatives of the Defense Production Administration and the Department of Defense, and such additional representatives of the Munitions Board and the Armed Services as may be agreed upon." Harry Ehle, International Resistance Co., and Don G. Mitchell, Sylvania Electric Products Inc., have been designated to represent the Army and Air Force, respectively, on the Board. The Navy will be represented by Vice Admiral A. G. Noble, Chief, Office of Naval Material, and the Munitions Board by Chairman John D. Small. W. W. Watts, assistant to DPA Administrator Harrison, has been serving as Acting Chairman of the Board pending the appointment of a permanent chairman. . . . The National Production Authority is considering a **method to maintain production of receiving tubes** at the highest possible level until defense requirements are known. This disclosure was made at a recent meeting of the Receiving Tube Industry Advisory Committee with NPA officials. Committee members advised NPA that the industry's biggest problem is obtaining adequate nickel supplies. NPA pointed out that because of the industry's importance to the defense effort, the agency plans to reserve a supply of nickel which would be made available to the industry during the second quarter of this year. . . . All manufacturers of radio and television receivers have received a **letter from RTMA General Manager James D. Secrest** informing them that a critical situation may exist on the availability of selenium suitable for use in selenium rectifiers. A questionnaire accompanying the letter called for each company's anticipated requirements for selenium rectifiers. When returned to RTMA Washington headquarters, the data will be compiled on an industry-wide basis and presented to the National Production Authority so that agency will know the over-all industry requirements for selenium. Data submitted by individual companies will be confidential, and will be identified by code number rather than by name. . . . The National Production Authority was warned that the electronic industry is running into trouble in obtaining a specific type of copper used in magnetrons, resistors, and other equipments. **The shortage of OFHC copper** is said to be due to a shortage of facilities to process and draw the metal as well as to the copper shortage. NPA has moved to aid the industry in obtaining the metal. At the same time it has been learned that the Electronic Production Board is acting now to prevent an expected acute shortage of tungsten in the radio-television industry.

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[1] The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of March 16, March 23, March 30, April 6, and April 13, published by the Radio-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

## Military News

The Armed Services plan to issue a **new policy on reservists** which may assist industry in planning with regard to valuable trained technical personnel who are members of the Reserve. One of the major objectives of the policy is to give reservists not yet called definite information as to whether they may be summoned to active duty, and when. Further, it is expected to provide for release of some reservists now in the service. . . . **Brig. Gen. George Irving Back, General MacArthur's Signal Officer in Tokyo since 1947, has been nominated by the President to be Chief Signal Officer of the U. S. Army.** If confirmed by the Senate, he will succeed Maj. Gen. S. B. Akin, who was Chief Signal Officer from April 1, 1947, until he retired from the Army on March 31 of this year. During World War II, from September, 1944, to November, 1945, General Back served in the Mediterranean Theater of Operations, as Deputy Chief Signal Officer of the Allied Force Headquarters, and as Chief Signal Officer of the Mediterranean Theater. He also held several important wartime positions in the Office of the Chief Signal Officer in Washington. . . . **Maj. Gen. Francis H. Lanahan, Commanding General of the Signal Corps Center, Fort Monmouth, N. J., has been assigned to Supreme Headquarters Allied Powers in Europe** and is expected to report in the near future, according to the Department of the Army. His successor will be announced at a later date.

## Television News

Initial steps toward **lifting the "freeze" on new television station construction** were taken recently by the Federal Communications Commission with the issuance of a "Third Notice of Further Proposed Rule Making." Public hearings begin on June 11. Principal provisions of the FCC's new allocation plan are: Sixty-five or 70 uhf channels are to be made available for television in addition to present 12 vhf channels, laying ground work for about 2,000 television stations ultimately; 52 uhf channels will be made available throughout the United States on an intermixture basis with vhf channels; increased power for existing vhf stations is provided; adjacent channel separation for vhf and uhf stations is reduced to 70 and 65 miles, respectively, while the minimum separation of stations on adjacent channels is 60 miles for vhf and 55 for uhf; 31 vhf stations now operating will be required to shift channels if the plan is adopted; provision is made for earmarking about 10 per cent of the channels, chiefly in the uhf band, or about 200 stations for noncommercial educational use; earlier lifting of vhf "freeze" in Alaska, the Hawaiian Islands, Puerto Rico and the Virgin Islands is proposed. While the Commission definitely has "broken the ice" of the "freeze" and has moved towards a partial lifting in the near future, it is most unlikely that the construction ban can be lifted before autumn. The allocation of either 65 or 70 uhf channels will depend on how the FCC rules on a common carrier request for five channels in the 470–500 mc band. If this proposal is rejected by the FCC, Channel No. 14 would begin at 470 mc; but, if it is granted, the uhf broadcasting channels would commence at 500 mc. The FCC gave no indication when this issue would be decided. In rejecting industry proposals to keep uhf and vhf stations separate, the FCC said it had decided "that the adoption of an assignment table based on nonintermixture constitutes a short-term view of the problem and is inadvisable." In proposing a reduction in adjacent channel separation, the FCC stated that problems which have been encountered have not been serious and can be solved to a very considerable extent by improvements in receiver design which are neither difficult nor costly. The Commission added that it is of "the opinion that these separations should be based upon receiver performance which may reasonably be expected of manufacturers and not on the characteristics of the poorer receivers." Recognized as a victory for Commissioner Frieda Hennock, the reservation of some 200 assignments for noncommercial education use drew supplemental opinions from Chairman Coy and Commissioner Hennock. Commissioners E. M. Webster, George E. Sterling, and Robert F. Jones dissented in part from the decision. . . . **The standard intermediate frequency of 41.25 mc** recommended by the RTMA engineering department was "taken into account" by the FCC in making the uhf allocations, the Commission said in an appendix to its proposed order, and added that the use of an intermediate frequency of 111 mc in television sets "is not feasible at this time." . . . **The frequencies of 31 vhf stations now broadcasting will be shifted** if the FCC's new allocation plan is adopted as proposed. Only three such shifts were proposed in the Commission's earlier allocation plan. "In preparing the 'table' of television allocations," the Commission stated it "proposes to alter existing television authorizations in 31 instances. The alterations with respect to Channel 9 in the city of Cleveland, Ohio, Channel 5 in the city of Syracuse, N. Y., and Channel 6 in the city of Rochester, N. Y., resulted from the Commission's efforts to arrive at an equitable distribution of television channels between the United States and the Dominion of Canada. The remaining 28 channel substitutions resulted from the Commission's efforts to reduce interference, make available a reasonable number of channels, and to effect the maximum utilization of vhf television channels in the United States." The stations which would be required to change frequencies are listed in Table I.

TABLE I

| Licensee or Permittee | City Affected | Present Channel Assignment | Proposed Channel Assignment |
|---|---|---|---|
| WOI-TV | Ames, Iowa | 4 | 5 |
| WSB-TV | Atlanta, Ga. | 8 | 11 |
| WBRC-TV | Birmingham, Ala. | 4 | 6 |
| WTTV | Bloomington, Ind. | 10 | 4 |
| WBKB | Chicago, Ill. | 4 | 2 |
| WLWT | Cincinnati, Ohio | 4 | 5 |
| WKRC-TV | Cincinnati, Ohio | 11 | 12 |
| WCPO-TV | Cincinnati, Ohio | 7 | 9 |
| WXEL | Cleveland, Ohio | 9 | 8 |
| WNBK | Cleveland, Ohio | 4 | 3 |
| WLWC | Columbus, Ohio | 3 | 4 |
| WLWD | Dayton, Ohio | 5 | 2 |
| WHIO-TV | Dayton, Ohio | 13 | 7 |
| WOC-TV | Davenport, Iowa | 5 | 6 |
| WLAV-TV | Grand Rapids, Mich. | 7 | 8 |
| WSAZ-TV | Huntington, W. Va. | 5 | 8 |
| WJAC-TV | Johnstown, Pa. | 13 | 6 |
| WGAL-TV | Lancaster, Pa. | 4 | 8 |
| WAVE-TV | Louisville, Ky. | 5 | 3 |
| WHAS-TV | Louisville, Ky. | 9 | 11 |
| WMCT | Memphis, Tenn. | 4 | 5 |
| WTMJ-TV | Milwaukee, Wis. | 3 | 4 |
| WNHC-TV | New Haven, Conn. | 6 | 8 |
| WTAR-TV | Norfolk, Va. | 4 | 10 |
| WKY-TV | Oklahoma City, Okla. | 4 | 7 |
| WDTV | Pittsburgh, Pa. | 3 | 2 |
| WJAR-TV | Providence, R.I. | 11 | 10 |
| WHAM-TV | Rochester, N.Y. | 6 | 5 |
| WRGB-TV | Schenectady, N.Y. | 4 | 6 |
| WSYR-TV | Syracuse, N.Y. | 5 | 3 |
| WDEL-TV | Wilmington, Del. | 7 | 12 |

# IRE People

E. U. Condon

**E. U. Condon** (M'42–SM'43), Director of the National Bureau of Standards, was recently elected honorary member of the Société Française de Physique, the French Physical Society. This is an honor which has been accorded only nine other scientists in the history of the society.

Born in Alamogordo, N. M., in 1902, Dr. Condon received the B.A. degree in 1924, and the Ph.D. in 1926, both from the University of California. The recipient of one of the Rockefeller Foundation's National Research Council Fellowships, he continued his studies on the postdoctoral level in Germany from 1926 to 1927. Upon his return to America, he began his teaching career as a lecturer in physics at Columbia University, a career which included appointments to the University of Minnesota, to Princeton, and to the University of Pittsburgh.

In 1937 Dr. Condon left the classroom for the industrial laboratory, having been appointed associate director of the Westinghouse Research Laboratories in East Pittsburgh, where he was instrumental in working on the only large atom smasher to be operated by industry.

Beginning with the fall of 1940, Dr. Condon devoted all of his time to various phases of research for military purposes, both with Westinghouse and with many government projects. He became a consultant to the National Defense Research Committee when that group was first established by the late President Roosevelt. In this capacity he helped organize the Radiation Laboratory, for which he wrote the basic text on the subject of microwaves, introduced the radar program to the Westinghouse Corporation, and organized the microwave electronics research divisions of the company. In 1941 he became a member of the National Defense Research Committee's Rocket Program group.

In the autumn of 1941, he was appointed to a subcommittee of the NDRC, charged with studying the feasibility of producing atomic bombs. The investigations of this committee resulted in the recommendation that the United States proceed with a large-scale atomic bomb project immediately.

In 1945 he resigned from Westinghouse to become Director of the National Bureau of Standards by Presidential appointment. At the same time, by a second Presidential appointment, he became a member of the National Advisory Committee for Aeronautics. He also served, during 1945 and 1946, as scientific adviser to the Special Committee on Atomic Energy of the United States Senate which, under the chairmanship of Senator Brien McMahon, reported on the bill establishing the Atomic Energy Commission. In the summer of 1946, Dr. Condon was appointed to the President's Evaluation Board for "Operation Crossroads." He also belongs to the scientific advisory board of the Brookhaven National Laboratory of the AEC.

Dr. Condon has written more than 60 research papers, most of which are "required reading" in nuclear physics, as well as a large number of less rigorously technical articles. He is the co-author of two books, "Quantum Mechanics," published in 1929, and the "Theory of Atomic Spectrum," which appeared in 1935.

In 1944 he was elected to the National Academy of Sciences, and in 1947 to the American Academy of Arts and Sciences. He was president of the American Physical Society during 1946, and is a member of the governing board of the American Institute of Physics. Among the other scientific organizations to which he belongs are the American Institute of Electrical Engineers, the Institute of Aeronautical Sciences, and the American Association for the Advancement of Science. He is also a member of Sigma Xi, Phi Beta Kappa, and Sigma Pi Sigma. For his scientific achievements, particularly his contributions during the war, the Junior Chamber of Commerce of Pittsburgh elected him "Man of the Year" for 1945.

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**Joseph C. Ferguson** (A'41–SM'46), a former Chairman of the Fort Wayne IRE Subsection, died suddenly some time ago, it has been learned.

Born in Beaumont, Texas, in 1900, Mr. Ferguson received the B.S. degree from Louisiana State College in 1930, and did graduate work at Harvard University. He began his career as a junior radio engineer at Westinghouse Electric; shortly thereafter he joined RCA as a radio-design engineer. Four years later he became a senior radio engineer at the Farnsworth Television and Radio Corporation, where he was placed in charge of television-broadcasting equipment and design. In 1942 he was promoted to the position of chief engineer of the electronics apparatus division, in charge of the development and design of electronic equipment for the armed services.

Mr. Ferguson was a member of the RTMA Television Transmitter Committee, and served as chairman of their subcommittee on television studio facilities from 1944 to 1947.

W. B. Whalley

**W. B. Whalley** (A'37), engineering specialist for Sylvania Electric Products Inc., Bayside, L. I., N. Y., was appointed adjunct professor in electrical engineering at the Polytechnic Institute of Brooklyn. The appointment was made in recognition of his professional standing and of his distinguished experience in the electrical and electronic fields.

Professor Whalley received the B.A.Sc. degree from the University of Toronto in 1932; he served on the University's department of electrical engineering for the next four years, during which time he received the M.A. Sc. degree. Shortly thereafter he became associated with the Radio Valve Company of Toronto, and with the RCA Manufacturing Company. In 1940 he began his work on radar development for the Canadian government which included the technical organization of government-owned cathode-ray tube plant in Toronto for the production of radar screen tubes.

From 1943 to 1947 he was engaged in research at the RCA Laboratories in Princeton, N. J., where he continued work in the radar field and in the development of transmitting tubes and circuits for television. He also had done notable prewar work to improve television techniques, particularly with camera pickup tubes, television-receiver picture tubes, advanced circuits, and tubes for television systems.

His most recent work, at the Sylvania Physics Laboratories, has been in the field of simplified television receivers. He has consistently predicted a trend toward receiver simplification to reduce the number of required parts, increase reliability, reduce size of cabinets, and lower cost to the consumer. To prove his contentions, he has been demonstrating a 16-tube television receiver in direct comparison with commercial sets that require from 20 to 30 tubes for comparable results.

Professor Whalley is a member of Sigma Xi, the American Association of University Professors, and the American Physical Society.

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**Edward G. Hall** (A'43–M'48) has been appointed engineering representative for the sales and engineering service department of the Lenkurt Electric Company in San Carlos, Calif. Prior to joining Lenkurt, Mr. Hall had been for several years general transmission engineer for the Indiana Associated Telephone Corporation in Lafayette, Ind. His engineering experience includes also several years' affiliation with the Wisconsin Bell Telephone Company, and the Wilcox Electric Company of Kansas City, Mo.

**W. A. Weiss** (S'40–A'41–M'44) has been appointed manager of the new Sylvania radio receiving tube plant in Burlington, Iowa, it was announced recently. It is expected that he will take up his new duties early next autumn, at which time the construction of the plant will have been completed.

W. A. Weiss

Mr. Weiss, who has been manager of the Sylvania receiving tube plant in Emporium, Pa., since 1947, joined the company in 1940 as a student engineer in Emporium. In 1942 he was named supervisor of quality control and served in this capacity until 1947, when he became quality control manager for the entire Sylvania radio tube division.

Born in Alton, Ill., in 1913, Mr. Weiss was graduated from Pennsylvania State College in 1941 with the B.S. in electrical engineering. Before attending college, he had worked for a time with the Philco Corporation in Philadelphia. He is a member of the American Institute of Electrical Engineers and of The American Management Association.

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**A. M. Zarem** (S'42–A'46), director of Stanford Research Institute's Los Angeles division, and chairman of its applied physics research department, has been named one of "America's Ten Outstanding Young Men of 1950" on the United States Junior Chamber of Commerce list. His nomination was sponsored by the Alhambra, Calif., Junior Chamber, and was supported by numerous West Coast and Midwest educators, scientists, and newspaper and business executives.

The citation was made "for new photographic techniques making possible the taking of one million photos a second for the purposes of research in ballistics and similar fields."

Dr. Zarem, a native of Chicago, has won several awards based on photographic techniques he developed. His most spectacular achievement is the Zarem camera, so named by the Navy. With the speed of 100,000,000 frames per second, and an effective exposure time as fast as one-hundredth of a millionth of a second, it is the fastest known to science. This work opened a whole new field in physics research, termed "microtime," which has become vital in research on explosions, ballistics, shaped charges, and cavitation.

In 1948, Dr. Zarem was selected "The Outstanding Young Electrical Engineer in the United States" by Eta Kappa Nu.

In his capacity of chief administrative and scientific representative of Stanford Research Institute in Southern California, Dr. Zarem has concerned himself directly with the major problems facing the future development of the region, including mass transportation, water resources, and air pollution. He suggested, organized, and directed the First National Air Pollution Symposium sponsored by SRI and held in Pasadena in November, 1949.

In the past he has supervised groups of physicists and engineers whose work was directly used in the development of the atomic bomb. His department at Stanford Research Institute is presently engaged in studies dealing with fundamental explosion phenomena and with certain aspects of the effects of atomic weapons.

❖

**Eginhard Dietze,** a retired electrical engineer, died at Bradenton, Fla., according to a recent announcement. He had been in failing health for several months.

Mr. Dietze had had wide experience in many phases of telephone, and in the past few years had been associated with the development of the new and improved telephone set recently announced by Bell Telephone Laboratories.

Born on August 27, 1891, in Berlin, Mr. Dietze received his early education in Dresden, and moved to the United States in 1915. He enrolled at the University of Michigan where, in 1917, he received the B.S. degree in electrical engineering.

He then became associated with the American Telephone and Telegraph Company, where he engaged in telephone circuit transmission studies. Two years later he joined the company's department of development and research, which became a part of Bell Telephone Laboratories in 1934.

During World War II, Mr. Dietze was on leave of absence to the National Defense Research Committee, where he served as technical director of a group working on sonar.

A former member of the Institute, he served on the Annual Review Committee, the Papers Procurement Committee, and the Standards Committee; he was Chairman of the Electroacoustics Committee from 1946 to 1948.

❖

## CORRECTION

An error was made in the Fellow citation of A. G. Kandoian on page 572 of the May, 1951, issue of the PROCEEDINGS. The correct version of the citation is given below:

"For his important contributions to the development of antennas and navigation systems."

**Emanuel R. Piore** (A'38, M'42, SM'43, F'50) was appointed Deputy and Chief Scientist of the Office of Naval Research, it was announced recently. Up to now he has held the title of Deputy for Natural Sciences.

E. R. Piore

Dr. Piore was graduated from the University of Wisconsin in 1930, and in 1936 received the Ph.D. from that university, where he remained until 1938 as an assistant instructor in physics.

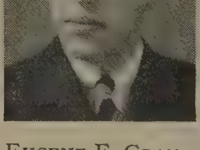
From 1938 to 1942 he was engineer in charge of the television laboratories of the Columbia Broadcasting System, and was instrumental in the development of color television. In 1944, from a position as senior physicist in the Navy's Bureau of Ships, he was called to active duty in the Navy Reserve as a lieutenant commander.

Dr. Piore joined the Office of Research and Inventions (which later became the Office of Naval Research) in 1946 as head of the electronics branch, and his service to ONR has been continuous except for the year 1948–1949, when he was a guest scientist in electronics at the research laboratory of the Massachusetts Institute of Technology.

A member of the Institute's Board of Directors, Dr. Piore belongs to Sigma Xi, the American Society of Naval Engineers, the Association for Computing Machinery, and is a Fellow of the American Physical Society.

❖

**Eugene F. Grant** (A'44), formerly chief of the computer branch, Air Force Cambridge Research Laboratories, has been appointed co-ordinator of electronics research and development at the W. L. Maxson Corporation in New York, N. Y.


Eugene F. Grant

Mr. Grant was born in Baker, Ore., and received the M.S. degree in electrical engineering from Oregon State College in 1942. He was then employed by the Westinghouse Research Laboratories, where he worked on the design of airborne radar. At the Sperry Gyroscope Company from 1945 to 1946, he was concerned with the design of Doppler radar systems. At the Air Force Cambridge Research Laboratories he was in charge of the design and construction of specialized digital computing machines.

Mr. Grant is a member of Phi Kappa Phi, Eta Kappa Nu, Sigma Pi Sigma, Pi Mu Epsilon, Sigma Tau, and Sigma Xi.

# Report of the Secretary—1950

TABLE I.—TOTAL MEMBERSHIP DISTRIBUTION BY GRADES

| Grade | As of Dec. 31, 1950 Number | As of Dec. 31, 1950 % of Total | As of Dec. 31, 1949 Number | As of Dec. 31, 1949 % of Total | As of Dec. 31, 1948 Number | As of Dec. 31, 1948 % of Total |
|---|---|---|---|---|---|---|
| Fellow | 323 | 1.1 | 284 | 1.1 | 259 | 1.1 |
| Senior Member | 2,726 | 9.4 | 2,421 | 9.0 | 2,192 | 9.4 |
| Member | 3,803 | 13.1 | 3,559 | 13.3 | 3,334 | 14.2 |
| Associate | 14,590‡ | 50.3 | 12,309† | 46.0 | 11,713* | 50.0 |
| Student | 7,560 | 26.1 | 8,196 | 30.6 | 5,939 | 25.3 |
| Totals | 29,002 | | 26,769 | | 23,437 | |

* Includes 1,260 Voting Associates.
† Includes 1,159 Voting Associates.
‡ Includes 1,072 Voting Associates.

TABLE II.—FIVE-YEAR ANALYSIS OF U. S. FOREIGN MEMBERSHIP

| | 1950 | 1949 | 1948 | 1947 | 1946 |
|---|---|---|---|---|---|
| TOTAL | 29,002 | 26,769 | 23,437 | 21,037 | 18,154 |
| U. S. & Possessions | 26,708 | 24,434 | 21,048 | 18,723 | 15,898 |
| Foreign (including Canada) | 2,294 | 2,335 | 2,389 | 2,314 | 2,256 |
| Per Cent Foreign | 7.9 | 8.72 | 10.2 | 11.0 | 12.4 |

TO THE BOARD OF DIRECTORS,
THE INSTITUTE OF RADIO ENGINEERS

Gentlemen:

The tremendously expanding technical field which the Institute represents is reflected in the activities of the organization as portrayed in the following pages of this Report of the Secretary for the year 1950. Not only is the factor of growth shown by the continued increases in almost every item which could represent an index, but also by the gradually evolving pattern of dealing with diversified membership interests by technical fields which the Professional Group method represents. This latter approach seems to be progressing successfully in the direction of providing enhanced value of membership to the individual in the Institute.

Attention is particularly directed to the figures concerning membership, Technical Committees and their meetings, and the financial statement.

The 1950 National Convention scored a record registration of 17,689 persons as compared with 15,710 in 1949, held 36 technical sessions with 170 papers as compared with 33 and 170, respectively, in 1949, and had 253 exhibits as compared with 225 in 1949. Other meetings were: Fourth Annual Spring Technical Conference (Sponsored by Cincinnati Section), 1950 IRE Technical Conference (Sponsored by Dayton Section), Radio Fall Meeting (Syracuse, N. Y.), IRE/URSI Meeting, Commissions 1, 6, 7, 4 (Washington, D. C.), Conference on Improved Quality Electronic Components (Washington, D. C.), IRE West Coast Convention (Long Beach, Calif.), New England Radio Engineering Meeting (Boston, Mass.) IRE Conference on Electron Devices (University of Michigan, Ann Arbor, Mich.), National Electronics Conference (Chicago, Ill.), IRE/AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine (New York, N. Y.), National Conference of the IRE Professional Group on Vehicular Communications (Detroit, Mich.), IRE/AIEE Conference on Electron Tubes for Computers (Washington, D. C.), and UHF-Microwave Conference (Sponsored by Kansas City Section).

The problem of the growing deficit because the revenues from Student dues have, in recent years, been insufficient to cover the basic cost of providing students with the PROCEEDINGS has been overcome by increasing them to $5.00 per year, which amount will become effective in 1951. There has been no increase in Student dues since the establishment of the Student grade in 1932.

The year 1951 brings new problems due to changed world conditions with the inevitable shiftings of industrial and economic emphasis, and technical personnel, that always occur during a National emergency. The management of the Institute has geared itself to keep track of these trends better to organize for providing a maximum of service to its members as changes take place.

Respectfully submitted,

Haraden Pratt

HARADEN PRATT, *Secretary*

February 23, 1951

## Membership

At the end of the year 1950, the membership of the Institute, including all grades, was 29,002, an increase of 2,233, or 8 per cent over the previous year. The 2,233-member increase in 1950 was less than 2,400 and 3,332, the increases for 1948 and 1949, respectively. The percentage increase was 11 per cent in 1948, and 14 per cent in 1949. The membership trend from 1912 to date is shown graphically in Fig. 2.

The distribution of members in the various grades for the years 1948, 1949, and 1950 is shown in the accompanying plot, Fig. 3. Actual figures for 1948, 1949, and 1950 are shown in Table I. Of the 13,518 nonvoting Associates, 3,919 have been in that grade for more than five years. 1,786 Students transferred to the Associate grade in 1950, compared to 951 in 1949. The membership ratio of Associates to higher grades was 6 to 1 in 1944, 4 to 1 in 1945, less than 3 to 1 in 1946, and about 2 to 1 in 1947, 1948, 1949, and 1950, a very satisfactory trend. Monetary exchange difficulties and devaluation of the pound have contributed to the reduced increase in foreign membership during the past three years, shown in Table II.

It is with deep regret that this office records the death of the following members of the Institute during the year 1950:

*Fellows*

Edwin H. Colpitts (A'14–F'26)
Lawrence C. F. Horle (A'14–M'23–F'25)
Karl G. Jansky (A'28–M'34–SM'43–F'47)
Fred Kolster (A'12–M'13–F'16)
William D. Loughlin (A'29–M'29–F'35)
Roger M. Wise (A'26–M'30–F'37)

*Senior Members*

John E. Allen (A'23–SM'45)
Francis E. Bash (SM'45)
John F. Bates (A'36–SM'45)
Norman R. Beers (SM'50)
Loyd A. Briggs (M'29–SM'43)
Roy W. Chesnut (M'42–SM'43)
Frederick A. Cobb (A'26–M'28–SM'43)
B. Ray Cummings (A'18–M'20–SM'43)
Hammond H. Hollis (A'28–M'38–SM'43)
James H. Ludwig (A'37–VA'39–SM'48)
George O. Milne (M'43–SM'43)
John V. Murphy (SM'47)
Nicholas M. Oboukhoff (M'23–SM'43)
Otis W. Pike (A'26–M'29–SM'43)
Earl H. Schoenfeld (A'35–SM'45)

*Members*

M. S. Alexander (A'43–M'44)
A. B. Ellis (A'35–M'45)
Margaret S. Heagy (M'46)
Leonard D. Scalise (M'49)
Paul E. Waples (M'46)
Robert C. Woodhead (A'37–M'46)

*Voting Associates*

George W. Bain (A'31–VA'39)
John C. Bain (A'36–VA'39)
Robert J. Hare (A'32–VA'39)
Aristote Mavrogenis (A'38–VA'39)
Albert J. Muchow (A'35–VA'39)
Walter E. Poor (A'29–VA'39)

*Associates*

W. B. R. Agnew (A'44)
Morris D. Douglas (A'48)
Gaylord E. Durham (A'44)
Elmer W. Everson (A'47)
Loren B. Harrell (A'43)
Werner C. Kruger (A'48)
Richard L. Laudenslager (A'49)
Douglas S. Mackiernan (A'41)
Arthur G. Mohaupt (A'44)
Charles E. Morris (S'45–A'48)
Alexander Nadosy (A'39)
Alfred P. O'Connor (A'42)

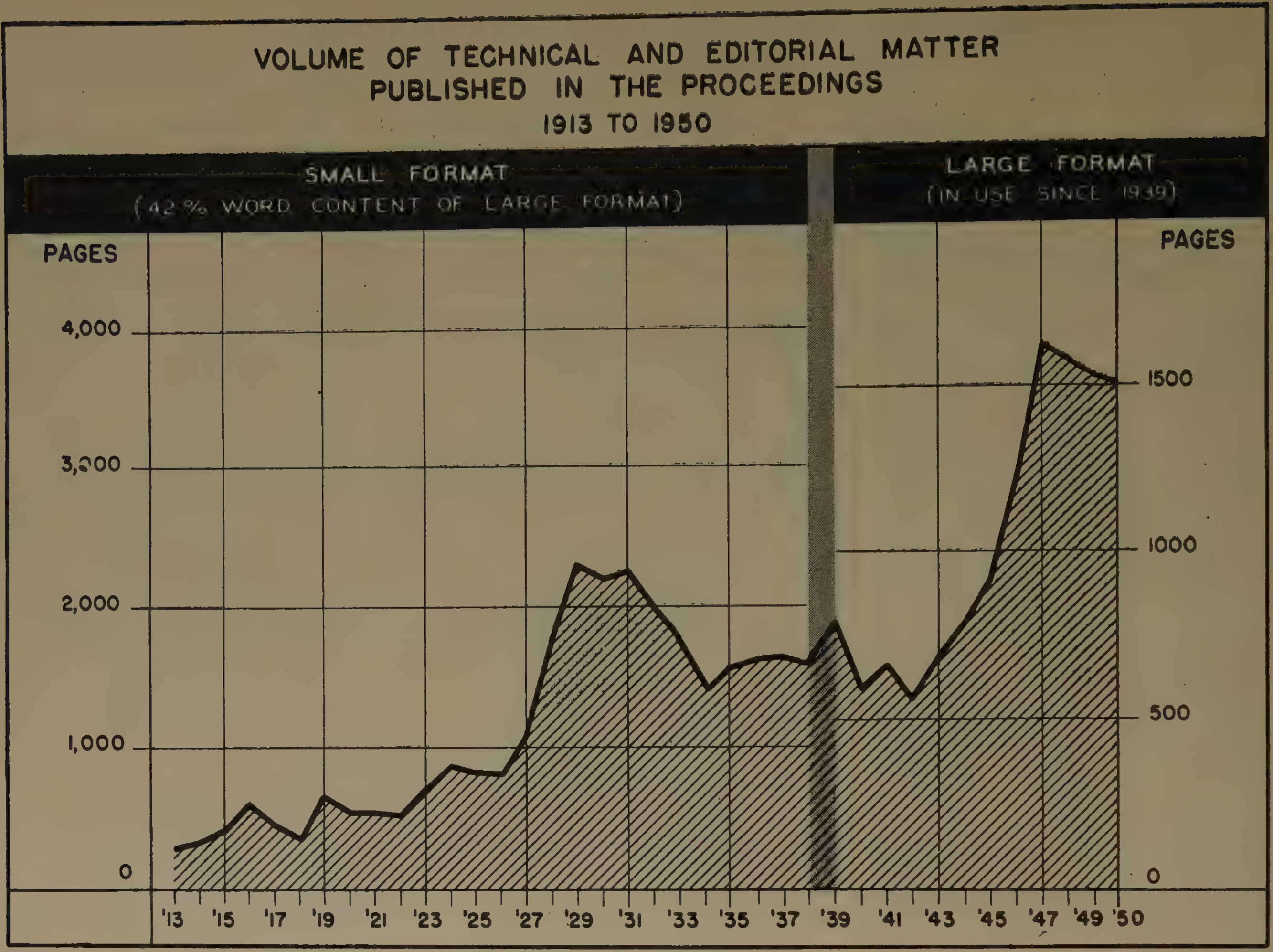
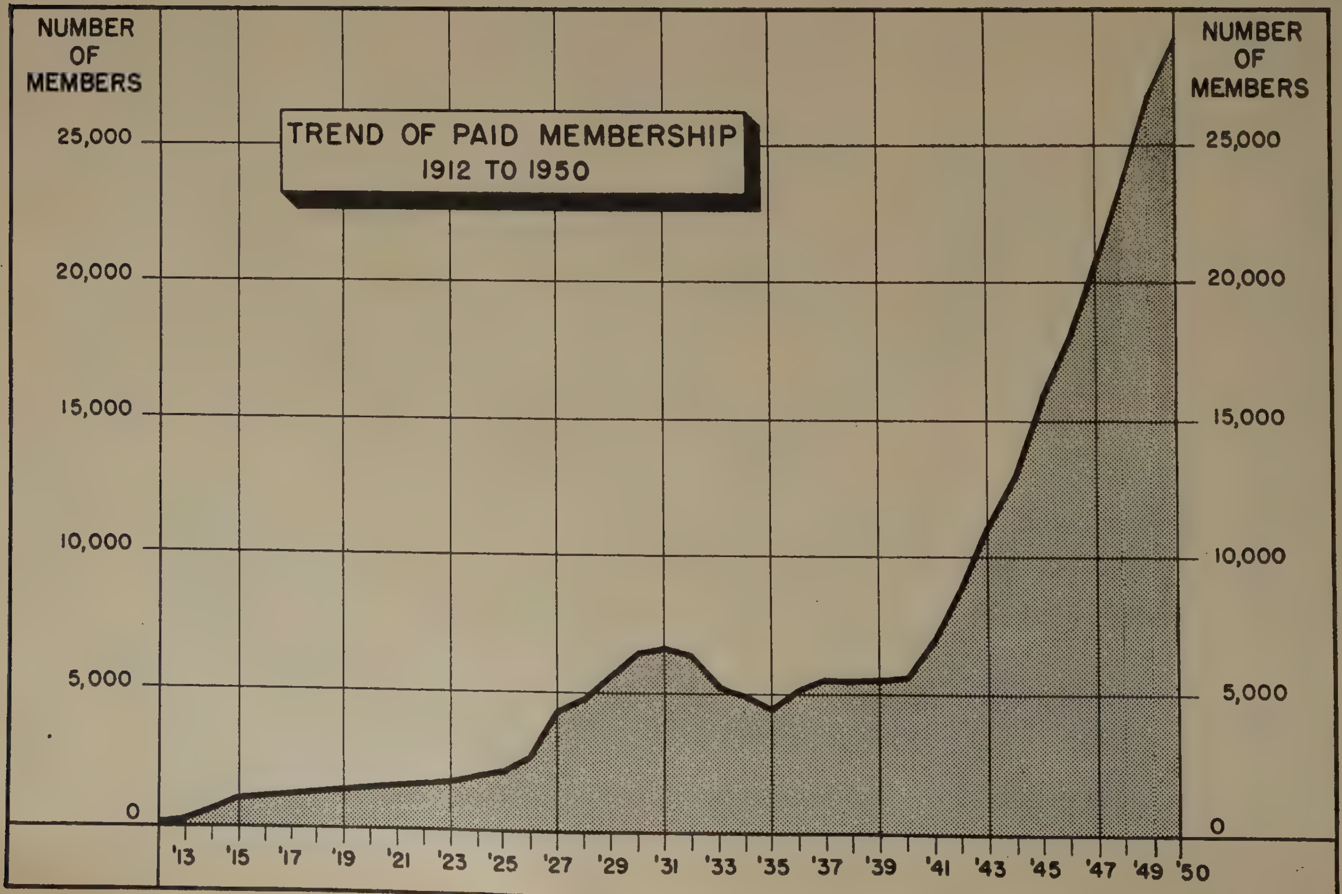

VOLUME OF TECHNICAL AND EDITORIAL MATTER PUBLISHED IN THE PROCEEDINGS 1913 TO 1950
SMALL FORMAT
(42% WORD CONTENT OF LARGE FORMAT)
LARGE FORMAT
(IN USE SINCE 1939)
PAGES
4,000
3,000
2,000
1,000
0
PAGES
1500
1000
500
0
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49 '50

Fig. 1


NUMBER OF MEMBERS
TREND OF PAID MEMBERSHIP 1912 TO 1950
25,000
20,000
15,000
10,000
5,000
0
NUMBER OF MEMBERS
25,000
20,000
15,000
10,000
5,000
0
'13 '15 '17 '19 '21 '23 '25 '27 '29 '31 '33 '35 '37 '39 '41 '43 '45 '47 '49 '50

Henry F. Parks (A'42)
S. Sampat (A'47)
Robert H. Schott (S'49–A'50)
William F. Seeman (A'35)
C. S. Sesha-iyer (A'49)
George S. Stancu (S'46–A'47)
Max D. Weinberg (A'48)

*Students*

William M. Cory (S'49)
Robert W. Gardner (S'49)
Donald B. Garrett (S'47)
Henry P. Kun (S'49)
Donald W. Maas (S'49)
Nicholas Sidjak (S'49)

## Fiscal

A condensed summary of income and expenses for 1950 is shown in Table III, and a balance sheet for 1950 is shown in Table IV. Income and expenses for each year since 1914 are shown graphically in Fig. 4.

TABLE III.—SUMMARY OF INCOME AND EXPENSES,

| | 1950 | |
|---|---|---|
| *Income* | | |
| Advertising | $211,021.40 | |
| Members Dues and Conventions | 434,596.55 | |
| Subscriptions | 50,575.45 | |
| Sales Items, Binders, Emblems, etc. | 15,811.38 | |
| Investments Income | 11,415.65 | |
| Miscellaneous Income | 282.35 | |
| TOTAL INCOME | | $ 723,702.78 |
| *Expenses* | | |
| PROCEEDINGS Editorial Pages | $170,806.71 | |
| Advertising Pages | 116,565.09 | |
| Yearbook | 49,030.14 | |
| Section and Student Branch Rebates | 28,916.62 | |
| Sales Items | 9,090.08 | |
| Miscellaneous Printing | 2,387.15 | |
| General Operations | 174,707.08 | |
| Convention Cost | 108,975.04 | |
| TOTAL EXPENSES | | $ 660,477.91 |
| *Surplus* | | $ 63,224.87 |
| Reserve for Depreciation | | 7,106.77 |
| NET SURPLUS | | $ 56,118.10 |

TABLE IV.—BALANCE SHEET—DECEMBER 31, 1950

| | | |
|---|---|---|
| *Assets* | | |
| Cash and Accounts Receivable | $234,505.04 | |
| Inventory | 13,267.59 | |
| Total Current Assets | | $ 247,772.63 |
| Investments at Cost | 488,059.14 | |
| Building and Land at Cost | 356,687.30 | |
| Furniture and Fixtures at cost | 76,014.32 | |
| Other Assets | 14,602.49 | |
| TOTAL | | 935,363.25 |
| TOTAL ASSETS | | $1,183,135.88 |
| *Liabilities and Surplus* | | |
| Accounts Payable | $ 8,891.00 | |
| Federal Taxes on Emblems, etc. | 85.58 | |
| Total Current Liabilities | | $ 8,976.58 |
| *Deferred Liabilities | | 260,103.96 |
| Total Liabilities | | $ 269,080.54 |
| Reserve for Depreciation | | 26,878.80 |
| Surplus—Donated | $597,281.03 | |
| Surplus—Earned | 289,895.51 | |
| TOTAL SURPLUS | | 887,176.54 |
| TOTAL LIABILITIES AND SURPLUS | | $1,183,135.88 |

* 1951 Items, PROCEEDINGS for members and subscribers, Advertising, and Convention Service.

## Editorial Department

During the year 1950 there were published in the PROCEEDINGS OF THE I.R.E. a total of 2,532 pages, including covers. Of these, 1,516 were editorial pages and 1,016 advertising pages.

The total of 2,532 pages published during the year compares with 2,404 in 1949, and 2,452 in 1948. The number of editorial pages (1,516) compares with 1,532 in 1949, and 1,592 in 1948. Advertising pages numbered 872 in 1949 and 860 in 1948. The number of editorial pages published each year since 1913 is shown in Fig. 1.

Technical papers totaling 186 were published in 1950, as against 183 in 1949. In addition, 6 IRE Standards were published in 1950 as against 4 in 1949. Authorship of these papers was by 252 individuals of whom 182, or 73 per cent, were members of the Institute. In 1949, 183 of the 252 authors, or 73 per cent, were members.

The volume of papers submitted for publication continues at a high rate. During 1949, 287 papers totaling an estimated 1,894 PROCEEDINGS pages were submitted, or an average of 24 papers and 158 pages per month. During 1950, 271 papers of 1,547.2 pages were received, or 23 papers and 129 pages per month.

The backlog of papers on hand in the Editorial Department at the end of 1950 consisted of 141 papers totaling 777 PROCEEDINGS pages, of which 81 papers or 436 pages had been accepted for publication, the remainder being under review. This represented a noticeable increase over the record low backlog total at the end of 1949 of 103 papers or 585 pages, but was substantially lower than the 1948 total of 141 papers or 981 pages.

In a move to consolidate the technical material in the PROCEEDINGS, the Waves and Electrons Section was discontinued with the January, 1950, issue. This did not affect the contents of the journal, but offered a more closely integrated arrangement of technical and editorial material.

An important innovation in papers procurement activities is being carried out by the Tutorial Papers Subcommittee of the IRE Committee on Education. The Subcommittee, under the chairmanship of Ernst Weber, is engaged in procuring, and recommending for publication, a series of tutorial papers on a wide variety of topics of both present and historical interest. These papers, which will appear at frequent intervals during 1951 and 1952, are to be educational in nature, of exceptional clarity, and prepared in each case by an authority in the corresponding field. It is believed that members in all grades will derive considerable benefit from this series.

During the year, the Editorial Department co-operated closely with IRE Professional Groups on the publication of technical papers and news items of interest to Group members in order to give this important Institute activity full expression in the pages of the PROCEEDINGS. Some of the steps taken by the Editorial Department are as follows: the procurement of papers of particular interest to Groups; the referral of submitted papers to Professional Groups for publication consideration; prompt publication of papers recommended by Groups; publication of 100-word abstracts of papers delivered at Group-sponsored conferences; publication of monthly columns on Professional Group News in the Institute News and Notes Section of the PROCEEDINGS; and bimonthly publication of a list of Professional Groups, together with the names and addresses of their chairmen. In addition, the Advertising Department frequently devoted the first page to publicizing Group-sponsored meetings, as well as other IRE conferences.

In recognition of the important work of the Editorial Administrative Committee in determining the acceptability of technical material for publication, the Board of Directors raised its status to that of a Standing Committee of the Institute and changed its name to Administrative Committee of the Board of Editors.

The 1950 IRE Directory (formerly known as the IRE Yearbook) was published on a considerably shortened production schedule, despite the substantial increase in size over prior years. The closing date for membership listings was changed from December 31 of the previous year to May 15 of the current year, and the mailing date was shifted from July to September 15.

The Editorial Department is directed by Editor Alfred N. Goldsmith in matters of editorial policy, content, and format, and by Executive Secretary George W. Bailey in matters of finance and administration. Both function through Technical Editor E. K. Gannett and an effective staff. It has been greatly assisted by the counsel and co-operation unstintingly given by the members of the Board of Editors, Papers Review Committee, and the Administrative Committee of the Board of Editors.

## Technical Activities

*Technical Committees.* During 1950, 23 Technical Committees, with their subcommittees, task groups and planning committees for conferences or symposia, held 146 meetings at the Institute Headquarters, five meetings at the Hotel Commodore during the National Convention, and ten meetings at other points, a grand total of 161 meetings.

A report on the activities of the IRE Technical Committees and their Subcommittees is compiled in the office of the Technical Secretary and published each month in the PROCEEDINGS.

In addition to IRE committee meetings, the National Television System Committee held 40 meetings at Headquarters. Several of the NTSC Panels requested and received secretarial assistance from the Office of the Technical Secretary. Eleven meetings of committees jointly sponsored by the IRE and other outside organizations were also held at Headquarters, many requiring secretarial services. During the year 21 conferences, symposia, technical sessions, and national meetings were sponsored by IRE Professional Groups, by Technical Committees or co-sponsored by the IRE and other organizations. All of these functions were well attended and met with enthusiastic audience response. The major portion of the work involved in the preparation of

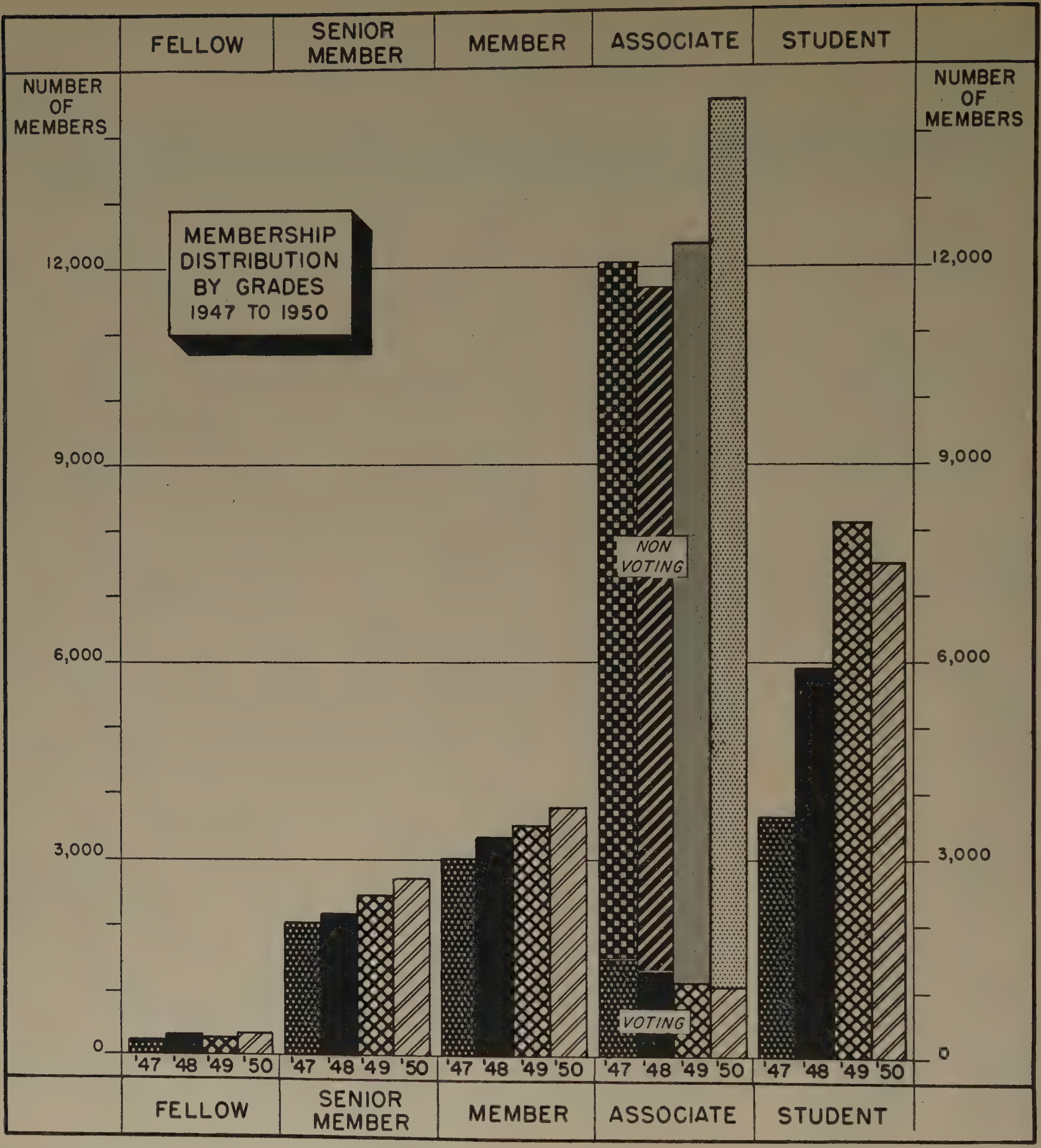


Fig. 3

these joint conferences and symposia was done by the Office of the Technical Secretary.

The policy of publishing Standards in the PROCEEDINGS was adopted late in 1949. Issues of the 1950 PROCEEDINGS containing Standards are identified by a red band on the cover and spine. This method of publication is still being carried on and reprints of Standards appearing in the PROCEEDINGS may be purchased from Headquarters for a nominal sum. Six standards were published during 1950.

The following standards were prepared by Technical Committees and appeared in various issues of the PROCEEDINGS during 1950:

*Standards on Electron Tubes: Definitions of Terms, 1950.*

*Standards on Electron Tubes: Methods of Testing, 1950.*

*Standards on Designations for Electrical Electronic and Mechanical Parts and their Symbols, 1949* (appeared Feb. '50).

*Standards of Television: Methods of Measurement of Television Signal Levels, Resolution, and Timing of Video Switching Systems, 1950.*

*Standards on Television: Methods of Measurement of Time of Rise, Pulse Width, and Pulse Timing of Video Pulses in Television, 1950.*

*Standards on Wave Propagation: Definitions of Terms, 1950.*

The following Standards, prepared by IRE Technical Committees, have been approved by the Standards Committee and will be published in forthcoming issues of the PROCEEDINGS.

*Standards on Circuits: Definitions of Terms in Network Topology, 1950.*

*Standards of Electroacoustics: Definitions of Terms, 1950.*

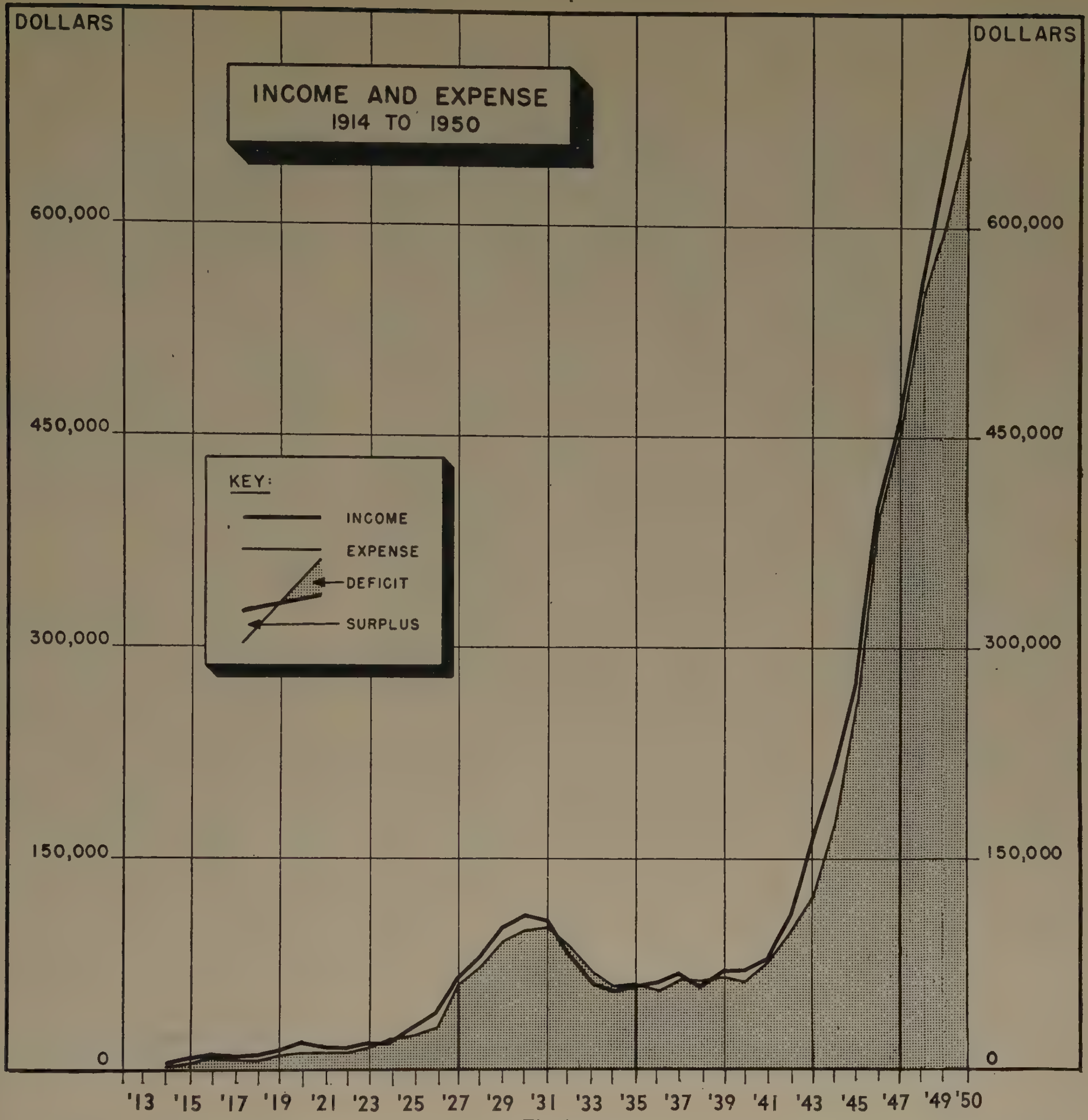


Fig. 4

*Standards on Electronic Computers: Definitions of Terms, 1950.*
*Standards of Abbreviations of Radio-Electronic Terms, 1950.*
*Standards on Television: Methods of Measurement of Electronically Regulated Power Supplies, 1950.*

The Annual Review Committee prepared its survey, "Radio Progress During 1950," which appeared in the April, PROCEEDINGS 1951, issue.

The Master Index of Terms, prepared in 1949 under the supervision of the Definitions Co-ordinating Subcommittee, was made available to all IRE Committees and, upon request, to other societies. It is presently being revised for the IRE Definitions Co-ordinating Subcommittee by the Office of the Technical Secretary.

The Institute approved the formation of the IRE Co-ordination Committee with International Technical Organizations. This Committee will provide liaison with CCIR and other international standardizing bodies with which the IRE is concerned. Axel G. Jensen has been appointed Chairman. This Committee will report directly to the IRE Executive Committee; it will work through the U. S. Department of State with the CCIR; and with other international bodies. When the CCIR or any other group requests IRE assistance, this new Committee will establish an ad hoc group to study the problem and provide a recommended solution. When the Standards Committee feels that a matter should be brought to the attention of an international standardizing body, it will in turn be referred to this Committee.

The Joint IRS Committee was formed by IRE, RTMA, and SMPTE to avoid duplication of effort in the preparation of standards in the fields associated with television. Axel G. Jensen, Millard W. Baldwin, Jr., and R. L. Garman are the IRE representatives on the Committee.

A Joint IRE/AIEE Committee on Noise Definitions has been organized for the specific task of critically reviewing all existing noise definitions and adding such new definitions as may be required.

The Institute has also approved the IRE participation in a newly organized committee which is known as the Joint IRE/AIEE Committee on High-Frequency Measurements. Ernst Weber, of the Polytechnic Institute of Brooklyn, is Chairman. The personnel of this Committee is composed of an equal number of IRE and AIEE members. E. I. Green is the leader of the IRE group. As a Joint Committee, it is autonomous and can engage in any activities which are of interest to Technical Committees of both sponsoring societies and to the IRE Professional Groups. It is the duty of

IRE members of this Committee to keep in contact with the IRE and AIEE activities. This Joint Committee has already sponsored two conferences.

The IRE maintains liaison with ASA Sectional Committees. During the year several IRE members were appointed to represent the Institute on various ASA Committees. The Office of the Technical Secretary is alert to the importance of maintaining adequate liaison with these committees.

*Professional Group System.* There are ten Professional Groups presently in existence in the following fields: Antennas and Propagation, Audio, Broadcast and Television Receivers, Broadcast Transmission Systems, Circuit Theory, Instrumentation, Nuclear Science, Quality Control, Radio Telemetry and Remote Control, and Vehicular Communications. The total membership is in excess of 8,500, and is increasing steadily. It is anticipated that a petition to form the 11th IRE Professional Group on Airborne Electronics will be approved by the Executive Committee. An example of the growth of the Professional Group System is evidenced by the fact that at the 1950 National Convention, five groups sponsored Symposia, while all of the Groups sponsored symposia at the 1951 National Convention. Each group was represented on the Technical Program Committee for the Convention.

IRE Sections are constantly informed of the activities of the Groups through the media of newsletters, conference notices and minutes from the office of the Technical Secretary. Steps are being taken in the various Sections towards the stimulation of Group activities. As a result of membership drives, many new members have been enrolled in IRE. During the year a number of Groups sponsored joint symposia, national meetings and technical sessions. All Groups are interested in securing papers for publication in the PROCEEDINGS. The Professional Group Manual was revised in June 1950.

*The Joint Technical Advisory Committee.* There were 11 meetings during the year. Members of the JTAC attended demonstrations given by CBS and the Hazeltine Electronics Corporation of various matters within their fields of interest. Volume V, Adjacent-Channel Interference In Monochrome Television, was published and submitted with a letter to the Federal Communications Commission; Volume VI, (correspondence and minutes) was authorized and work commenced; the publication of a supplement to Volume IV, Comments On The Proposed Allocation of Television Broadcast Services, was prepared. A 'Supplemental Statement' prepared for the FCC in connection with Volume IV was formally presented in evidence before the Commission. The membership was appointed for the year July 1, 1950 to June 30, 1951, and J. V. L. Hogan was unanimously elected Chairman. Melville Eastham and E. K. Jett were unable to continue their duties with JTAC, and were succeeded by I. J. Kaar and T. T. Goldsmith. Several reports from technical associations and industry have been circulated for comment, and reports on current topics exchanged with other technical bodies, such as the BBC.

## Section Activities

We were glad to welcome four new Sections into the Institute during the past year. They are as follows:

| | |
|---|---|
| Evansville-Owensboro | (March) 1950 |
| Hawaii | (February) 1950 |
| Miami | (February) 1950 |
| Vancouver | (September) 1950 |

The total number of Sections is now 57. There has been a substantial increase in membership of these Sections with a few exceptions. It should be borne in mind that most Sections with noticeable decreases in membership released substantial numbers of members to new Sections.

The Subsections of Sections now total 12, the following being formed in 1950:

Binghamton (Syracuse Section)
  May, 1950
Mid-Hudson (New York Section)
  November, 1950

## Student Branches

The Institute's program with respect to Student Branches continued to flourish in 1950. The number of Student Branches formed during 1950 was 17, 14 of which operate as joint IRE/AIEE Branches. The total number of Student Branches is now 105, 60 of which operate as joint IRE/AIEE Branches. This increase of Student interest was accompanied by a large increase in Student members.

Following is a list of the Student Branches formed during 1950:

University of Akron, University of British Columbia, Polytechnic Institute of Brooklyn, Bucknell University, University of Connecticut, Cooper Union, University of Denver, Drexel Institute of Technology, Louisiana State University, University of Miami, Michigan College of Mining and Technology, New Mexico College of Agriculture and Mechanic Arts, Oklahoma Agricultural and Mechanical College, Rensselaer Polytechnic Institute, Stevens Institute of Technology, and University of Toronto.

# Books

### Radio Laboratory Handbook, Fifth Edition by M. G. Scroggie

Published (1950) by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London S.E. 1, England. 409 pages+5-page index+14-page appendix. 215 figures. 7×4½. 15s.

This is one of a series of radio books issued by the magazine *Wireless World*, the first of which appeared in 1938. The present fifth edition of this work, published in 1950, has been thoroughly brought up to date.

It is not a textbook on radio theory, but, as its name indicates, a reference book of material useful for workers in radio, designed to be useful both to the amateur enthusiast and to the professional radio engineer.

The material treated is grouped under the main headings of sources of power and signals, indicators, standards, laboratory equipment in general, measurements of resistance, inductance and capacitance, and tests of amplifiers and receivers. A new chapter is added on very-high-frequency work, with special reference to its relation to lower-frequency procedures. A reference chapter compiling useful fundamental formulas and numerical constants is a convenient feature.

With such a wide coverage of subject matter, exhaustive treatment of any subject is, of course, out of the question. A practical working knowledge of fundamental circuit theory and of elementary vacuum-tube functions on the part of the reader is assumed. Working diagrams of special circuits and arrangements are clearly given and discussed, but for further study by those interested, reference is given to the original published articles.

Space is given to sensible practical advice on the organization of an experimental laboratory, the choice of necessary apparatus, and instructions for the building of certain devices for measurement.

The style of presentation is, throughout, easy and pleasing, the choice of type size is good, and, in spite of the fact that the volume is of pocket size, the diagrams and illustrations do not suffer.

The amateur should find this a most helpful guide, and the professional engineer a handy and comprehensive reference book.

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### TV Master Antenna Systems, by Ira Kamen and Richard H. Dorf.

Published (1951) by John F. Rider Publishers, Inc., 480 Canal St., New York 13, N. Y. 352 pages+4-page index+vii pages. 234 figures. 5⅝×8⅝. $5.00.

This book was designed to be a working manual for many kinds of organizations and people concerned with master-television systems. The authors have succeeded to a remarkable degree in satisfying technical completeness and yet insuring the layman's comprehension by treating problems and their solutions. Although primarily presented for those whose interests are in manufacture, sales, installation, maintenance, and use, this is also a tutorial work of interest to radio engineers.

The reader is introduced to the subject by a discussion of basic television antenna systems. Descriptions of four nonamplified and ten amplified master-antenna systems follow. All but two of the systems discussed are replete with circuit diagrams giving component values. The two exceptions, Lynmar and Multitenna, give the dubious and provocative excuse that for "patent reasons" the information is not available. The authors may be commended for their adequate presentation in the absence of co-operation from these manufacturers.

The table of contents and the index appear to be ample for a work of this kind. The information is clearly presented in an easily readable manner. The book also covers video distribution systems, and provides an appendix with landlord agreement, survey report, and tenant letter forms.

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### Radiation Monitoring in Atomic Defense by Dwight E. Gray and John H. Martens

Published (1951) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. 111 pages+2-page index+19-page appendix+iv pages. 20 figures. 5½×8. $2.00.

Were the title of this little book to read simply "Radiation Monitoring," the reviewer would commend both the authors and the publishers for producing a manual useful to the uninitiated who are confronted with the problem of monitoring the radioactive hazard of a laboratory, an X-ray installation, or a similar radiation source. But the title contains the additional words "In Atomic Defense." Herein lies the fault—not in the construction of the book, but in its objective.

By aiming their book specifically at civil defense, the authors tend to dignify the status of the radiation hazard attending an atomic bomb explosion. In this respect they are not to blame, for they are merely "following the party line" established by the many medical men who have been responsible for civil defense planning. By and large, defense planners have gotten the radioactive menace of the atomic bomb completely out of focus, until it has assumed proportions overshadowing the more important blast and fire hazards of the new weapon. Many local communities have set up civil defense plans patterned to a radiological monitoring procedure which is not only disproportionate to the hazard, but may actually be dangerous to the overall success of civil defense. Apparently the thought has been that if you have a Geiger counter, you have the panacea for atomic defense.

Atomic bombs are normally exploded above the surface of the earth in order to achieve maximum effectiveness. American experience with 12 A-bombs so exploded shows conclusively that with an above-surface burst the residual radiation hazard is very small, and negligible with a high air burst. The very violence of the atomic explosion sweeps all the radioactive fragments upward and sucks them into the stratosphere, where they disperse harmlessly.

If one thinks of the book purely as a manual for radiation monitoring without reference to atomic disaster, then one is quite favorably impressed that the writers have done a creditable job. Treating the subject in two parts, they first lay out the fundamentals of atomic energy and nuclear radiation. Then they point their remarks at individual instruments (cookbook fashion as they, themselves, admit), and describe how the basic types of Geiger counter, ion chambers, and similar instruments should be used. As a matter of style, the reviewer wishes that they had been consistent in using the capitalized form of Geiger counter; they use both, but mainly the "geiger" form appears in the text.

Factually, the book appears to be quite accurate with only occasional errata cropping up. For example, the authors state: "radiation remains a serious cause of injury at distances up to 1½ to 2 miles from ground zero." This is a misinterpretation of data published by the Atomic Energy Commission, and the figures should read "up to 1 mile from ground zero."

Both the authors are on the staff of the Atomic Energy Commission. This very fact will add weight to the "official nature" of the publication. It also tends to explain the nature of the book, for there are a number of AEC officials who have been responsible for the delusion that the Geiger counter is the "open sesame" to the realms of civil defense.

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### Industrial Instrumentation by Donald P. Eckman

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 353 pages+9-page index+31-page appendix+vi pages. 247 figures. 5½×8½. $5.00.

This up-to-date textbook on industrial instrumentation is intended for the undergraduate student in engineering. The author assumes that the student has an acquaintance with the elements of calculus, mechanics, and thermodynamics, as well as a general course in physics.

As an introduction to the body of the text, the initial chapter covers the qualities of measurements, defines terms, and discusses the response of first- and second-order physical systems. Approximately one third of the text is devoted to temperature measurements (4 chapters). Then a chapter each is devoted to composition analysis, mechanical measurements, pressure and vacuum measurements, head and liquid level, and flowmetering. The final chapter is devoted to process instrumentation.

In general, the various types of measuring instruments are adequately discussed. Since the description is supplemented by a discussion of the basis for the measurement, some of the limitations, as well as the range of the usually available industrial instruments, are included. For many types of instruments, e.g., potentiometers and optical pyrometers, the various forms made by different instrument manufacturers are discussed.

However, because of the large number of different types of instruments discussed, there are only brief comments on many forms, such as emission spectroscopy and mass spectroscopy.

Among some of the errors noted is the implication that a second-order system composed only of resistance and condensors will oscillate when a unit step function impulse is applied. Another is the statement that the volume change of a gas due to change in temperature is relatively small. Actually, the quantity involved is the pressure developed by an enclosed fluid.

However, these errors should not detract from the usefulness of the book as a text for the undergraduate. The material is easily readable. Numerous good illustrations and the many examples which have been worked out should give the student a clear picture of the field of industrial instrumentation. The sets of problems at the end of each chapter and the many tables in the appendix should also prove helpful. In general, very little mathematics is needed, since most of the equations are stated rather than derived, and the application of these equations, which is well illustrated, is not difficult.

Thus the book fulfills the undergraduate engineering student's need for a textbook in industrial instrumentation. It may also prove useful to some engineers who are in daily contact with instruments, but who need only a precursory knowledge of the fundamental principles of their operation.

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### The Principles of Cloud-Chamber Techniques by J. G. Wilson

Published (1951) by the Cambridge University Press, 51 Madison Ave., New York 10, N. Y. 127 pages+2-page index+2-page references+viii pages. 33 figures. 5½×8½. $2.75.

One of the Cambridge Monographs on Physics, this book is addressed to nuclear physicists who use or plan to use cloud chambers, or who may find an account of the cloud chamber as an instrument of precise measurement helpful.

The purpose of the cloud chamber is to study the motion of ionizing particles from photographs of the drops condensed on ions formed along particle trajectories. The early chapters of the monograph deal with the condensation process, ionization, and the general problems of the establishment of supersaturation and its persistence, on which the technique depends. Arrangements for photographing are then discussed and methods of measurement are described in detail, with particular reference to cosmic-ray investigations. A final chapter deals with the interpretation of photographs, a topic of increasing importance. Detailed specifications for the construction of cloud chambers are not included, but the advantages and disadvantages of various constructional materials are assessed in the light of experience.

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# URSI—IRE Spring Meeting*

WASHINGTON, D.C.,——APRIL 16–18, 1951

## SUMMARIES OF TECHNICAL PAPERS

### 1. RECENT MATHEMATICAL DEVELOPMENTS IN THE THEORY OF TROPOSPHERIC PROPAGATION

BERNARD FRIEDMAN

(New York University, New York, N. Y.)

The theory of tropospheric propagation fits naturally into the general theory of propagation in a stratified medium. It proves to be unnecessary to assume a flat earth and a modified index of refraction; the theory can be developed using a spherical earth and the ordinary index of refraction. The usual WKB treatment of the eigenvalue equation is extended and it is shown that propagation in an atmosphere without ducts can be reduced to propagation in a constant atmosphere by a suitable change of scale. The solution is approximated for an atmosphere with ducts, by means of Hankel functions of order one-fourth instead of the customary Hankel function of order one-third.

### 2. MARTYN'S THEORY OF MAGNETIC STORMS AND AURORAS

H. G. BOOKER

(Cornell University, Ithaca, N. Y.)

The theory supposes a neutral stream of charged particles from the sun as the origin of the ring current in magnetic storms. From this ring current, particles enter the earth's atmosphere along the auroral zone acquiring some ten times the energy of the primary particles. Visual auroras are caused by these secondary particles and drift of ionization from the auroral zones produce the current systems and the diurnal storm variation of the earth's magnetic field. Certain features of ionospheric observations are explained in terms of this current system.

### 3. IMPEDANCE MEASUREMENTS IN THE 50 TO 1,000 MEGACYCLE RANGE

F. J. GAFFNEY

(Polytechnic Research & Development Co., Inc., Brooklyn, N. Y.)

Several attacks have been made on the difficult problem of designing apparatus for impedance measurements in the 50–1,000 mc frequency range. These methods use an adaptation of slotted line techniques or the properties of tee junctions. The uhf equivalent of a low-frequency bridge is also described employing a novel variable-resistance element. Accuracies of the order of 5 per cent are obtained and possibilities of development for even higher precision are pointed out.

### 4. AIR-TO-AIR TROPOSPHERIC RADIO PROPAGATION

G. B. FANNING AND F. P. MILLER

(Wright-Patterson Air Force Base, Dayton, Ohio)

Recent flight tests performed over the Midwest by Aircraft Radiation Laboratory, Engineering Division, AMC, Wright-Patterson Air Force Base are described. Significant features of instrumentation and extensive experimental data for 250 mc and 3,000 mc are presented. The data indicate that many rapid and deep fades in signal level occur at ranges far short of the total horizon range. It is further concluded that significant observed field strength variations do not require severe associated meteorological anomalies for their explanation.

### 5. THE EFFECT ON PROPAGATION OF AN ELEVATED ATMOSPHERIC LAYER OF NONSTANDARD RETRACTIVE INDEX

L. H. DOHERTY

(Cornell University, Ithaca, N. Y.)

Elevated layers with nonstandard gradients of refractive index may radically affect the field of a transmitter above the layer. Serious difficulty in communication between aircraft flying above the layer may result from slight deviations from the standard atmosphere which frequently occur. Using a trilinear refractive index model of the atmosphere, it is shown that a decrease of field strength above the layer results within a certain range of distances. The interference pattern in a region of interference rays following this "radio hole" is shown. The effect of some nonlinear refractive index models is indicated. Comparisons are made with experimental results.

### 6. EXPERIMENTAL DISCRIMINATION OF THE FACTORS IN VHF RADIO WAVE PROPAGATION

A. W. STRAITON AND C. W. TOLBERT

(University of Texas, Austin, Tex.)

Results are reported for recent experiments in determining the relative contribution of the various possible propagation mechanisms affecting vhf radio field strengths at distances well beyond the optical horizon. Height-gain curves and their interpretation are given for distances up to 175 miles. Results of measurements of cross polarization are given providing additional information as to the relative contribution of the various propagation factors. Means of obtaining the scale of turbulence as used in the Booker-Gordon scattering theory are indicated and several measured values of this factor are given.

### 7. SOME CHARACTERISTICS OF 92.9-MC PROPAGATION OBSERVED AT A DISTANCE OF 127 MILES

R. J. WAGNER, JR.

(RCA Laboratories Division, Princeton, N. J.)

Vhf field strength observations of a distance of 127 miles from Boston to Riverhead are reported. For periods of weakest field strength the range of vertical angles of arrival extended up to about 26° while for stronger fields it extended only to 7° or less. During periods of rapid fading, a pronounced height diversity effect was found with field strength values ranging from 44 to 74 db below free space. When rapid fading was absent this range was 14 to 42 db below free space.

### 8. VALIDITY AND LIMITATIONS OF THE VAN VLECK-WEISSKOPF EQUATION FOR ATMOSPHERIC MICROWAVE ABSORPTION

T. F. ROGERS

(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

A detailed study is made of the Van Vleck-Weisskopf absorption equation for a collision-broadened line. Evidence indicates the equation to be essentially correct in the intermediate pressure region of $10^2$ to $10^{-2}$ mm of Hg. Attenuation values remote from resonance are studied and the frequency dependence of absorption discussed as a function of temperature and pressure. Below $10^{-3}$ mm Doppler broadening becomes the limiting line-width factor and line saturation effects occur for high incident power densities. Effects which become important at about one atmosphere, such as a shift in the resonance maximum and variations of line widths and structure with large partial pressure changes, are also studied.

### 9. ON THE DEFINITION OF VIRTUAL HEIGHT

JERRY SHMOYS

(New York University, New York, N. Y.)

The difficulties with the geometrical optics ("classical") definition of virtual height are discussed. The more general definition of virtual height in terms of the derivative of the phase of the reflection coefficient with respect to frequency is derived. It is then shown that the former definition can be derived from the latter, if the phase integral method is used. The two definitions are compared in the specific examples of linear, rectangular, Epstein, and parabolic charge distributions.

It is demonstrated, by means of examples, that the relation between virtual height and frequency derivative of phase is not valid when the reflected wave contains more than one pulse. In this case the frequency derivative of phase cannot be interpreted as the time delay of any one of the pulses.

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* Co-sponsored by the IRE Professional Group on Antennas and Propagation and the USA National Committee of URSI.

### 10. DISPERSION OF $F_2$-LAYER CRITICAL FREQUENCIES

M. L. Phillips and H. S. Moore

(National Bureau of Standards, Washington, D. C.)

Variations in dispersion of $f^\circ F_2$ about the monthly mean were studied with respect to solar activity, season, time of day, and geographical location. In general, the dispersion bears no simple relationship to the monthly mean $f^\circ F_2$. Comparison of the standard deviations for the five hours, 11 through 15, and 23 through 03, local standard time, individually, in summation, and in means of the five hours for each day, at three representative stations, showed a pronounced lack of independence in these data, especially for daytime hours. Evidence was found for consistent bias in the data obtained from certain stations. Nomograms were devised for applying the results of this analysis to practical transmission problems.

### 11. FLUCTUATIONS OF $F_2$ REGION BETWEEN STATIONS SEPARATED BY 100 TO 150 MILES

H. W. Wells

(Carnegie Institution of Washington, Washington, D. C.)

Analysis of ionosphere measurements made at three independent stations within a small geographical area revealed an occasional development of large differences in $F_2$-layer critical frequencies. Distances involved ranged from 100 to 150 miles. Mutually consistent values were observed during quiet periods by critical frequency differences in excess of 0.5 mc and on some occasions as great as 1.0 mc are observed. Cloud-like surges occurred independently at each station suggesting small physical dimensions of such disturbances. Results indicate the importance of transport by winds or redistribution by compressional-type traveling disturbances as affecting the instantaneous value of ionization density.

### 12. ANGLE OF ARRIVAL AND POLARIZATION AT FORT CHIMO

J. E. Hogarth

(Canadian Defence Research Board, Ottawa, Ont., Canada)

The direction of arrival of vertical incidence ionospheric echoes was measured by an interference method using four vertical-loop antennas placed at the corner of its horizontal square. The state of polarization was determined by phase measurements on a pair of mutually perpendicular loops. The measurements were designed to investigate spread echoes, echoes whose equivalent height changes rapidly, and echoes at a frequency higher than a critical frequency of the layer to which they appear to belong. Particularly important are polarization measurements on the $z$ trace, which is uncommon in the latitude of Fort Chimo.

### 13. USE OF LONG-DISTANCE BACKSCATTER TO DETERMINE SKIP DISTANCE AND MAXIMUM USABLE FREQUENCY

William Abel

(Raytheon Manufacturing Company, Waltham, Mass.)

Measurements of delay time in the 5- to 30-mc band between the leading edge of the backscatter pattern and the response of a beacon transponder has shown that delay time to the leading edge of the backscatter is a measure of the skip distance. Since the frequency on which backscatter is obtained is the maximum usable frequency (muf) for the skip distance of this frequency, the muf for other distances can be obtained by conventional methods. Backscatter can be obtained using peak-power outputs as low as 200–300 w. Methods are being developed to permit instantaneous determination of skip distance using ordinary transmitters and receivers together with suitable display equipment.

### 14. APPLICATION OF PUNCH CARDS TO THE ANALYSIS OF MULTIPATH IONOSPHERIC PULSE PROPAGATION RECORDS

W. C. Moore

(Boston University, Boston, Mass.)

The film records obtained from multipath ionosphere pulse propagation experiments made aboard high altitude missiles must be analyzed almost frame by frame because the variable parameter *altitude* has a different value for each frame. As many as 36 individual items of information may be of interest on a single frame, or may be necessary for comparison between frames selected from different portions of the record. This mass of data can be recorded as pencil marks on *mark-sense* cards which are then automatically punched and used in machines for all subsequent operations in the evaluation of the results of the experiment. Methods and operations are given for reduction of the data and punch-card analysis.

### 15. LOW-FREQUENCY ANTENNAS

P. S. Carter

(RCA Laboratories Division, Rocky Point, N. Y.)

For transmitting antennas in the frequency range from 15 to 300 kc, radiation efficiency is of primary importance. For ground-based receiving antennas the problem is different. Because of the high level of atmospheric noise, signal-to-noise ratio is little affected by efficiency and a receiving antenna of low efficiency and moderate directive gain is superior to a highly efficient omnidirectional radiator. Improvement in efficiency by top loading and reduction of ground losses are discussed. Several examples are given including the characteristics of multiple-tuned antennas. Improvement of signal-to-noise ratios by the Beverage wave antenna as used at Riverhead is described.

### 16. ANTENNA SYSTEMS FOR RADIO DIRECTION FINDING

E. C. Jordan

(University of Illinois, Urbana, Ill.)

A brief summary is made of antenna systems useful for radio direction finding. The characteristics of systems both small and large in wavelengths are considered. Methods of antenna synthesis are described that will yield narrow beam patterns and optimum patterns for linear and circular arrays. Problems peculiar to radio-direction finding antennas in the low-frequency, medium-frequency, and high-frequency bands are reviewed and some of the solutions obtained are indicated.

### 17. ELECTRICALLY SMALL ANTENNAS AND THE LOW-FREQUENCY AIRBORNE ANTENNA PROBLEM

J. T. Bolljahn

(Stanford Research Institute, Stanford, Calif.)

Properties of antennas which are small relative to their operating wavelength are discussed. Such antennas have radiation patterns corresponding either to that of a short dipole or to that of a small loop. The problem of a small antenna of arbitrary shape is treated in detail and a sensitivity parameter based upon the equivalent generator circuit is defined. An experimental procedure for determining the sensitivity parameters and radiation patterns of low-frequency aircraft antennas using scale models immersed in an electrostatic field is described, and several experimental results are shown. The problem of small loop-type antennas is discussed in a less detailed fashion, and some calculations of the bearing errors indicated by a low-frequency direction-finder loop in the vicinity of an idealized aircraft structure are presented.

### 18. ANTENNAS IN CONDUCTING MEDIA

R. K. Moore

(Cornell University, Ithaca, N. Y.)

A chief difficulty arises from high losses in the immediate vicinity of the antennas. The input impedance to an antenna in a conducting medium is essentially due to current flowing near the antenna and bears little relation to the field at a distance. Methods for analyzing certain kinds of antennas are shown and results of analysis are compared.

### 19. WAVE PROPAGATION OVER ROUGH SURFACES

W. S. Ament

(Naval Research Laboratory, Washington, D. C.)

Effect of ocean waves on the propagation of microwaves is discussed. The reflection coefficient for plane waves almost normally incident is derived on the assumption that all slopes of the ocean surface are nearly horizontal and that no portion of the surface is shadowed. Reflection formulas are given for near grazing incidence where the latter assumption is invalid. These are used in explaining the observed attenuation rates of theoretically completely trapped 3-cm normal modes. Alternative mechanisms and formulations of the problem are discussed.

### 20. SIMULTANEOUS MOBILE MEASUREMENT OF THE FIELD STRENGTHS OF TWO VHF RADIO STATIONS OVER IRREGULAR TERRAIN

R. S. Kirby
(National Bureau of Standards, Washington, D. C.)

Mobile field strength measurements at vhf to evaluate the random variations due to buildings and rough terrain are described. Measurement techniques and analysis of the data are discussed, leading to information on the distribution of median values of field strength within given sectors and the correlation between variations in two received fields as observed simultaneously. Overwater measurements show the effect of smooth terrain near the received location in that relatively minor variations of field strength were observed.

### 21. SUPPRESSION OF WAVES BY ZONAL SCREENS

H. E. Bussey
(National Bureau of Standards, Washington, D. C.)

Novel diffraction screens covering half a Fresnel zone are described. These may produce a complete null of field at a point and a weak field in the surrounding shadow area, thereby providing a method for greatly reducing reception of either a direct or a reflected wave at a fixed antenna. Application of this technique to reduction of the ground reflection wave which may cause interference fading in, for example, microwave radio relays, is discussed. An experimental test over an 800-foot path using microwaves is described.

### 22. THE EFFECT OF LOW-LEVEL ATMOSPHERIC CONDITIONS ON OVERWATER INTERFERENCE PATTERNS AT MICROWAVE FREQUENCIES

V. R. Widerquist and J. E. Boyd
(Georgia Institute of Technology, Atlanta, Ga.)

This paper gives a simple theoretical study of the effect of low-level atmospheric conditions on overwater interference patterns for microwave frequencies. Experimental data on radar interference patterns are given. Shorter range patterns reveal evidence of slight compression of the surface lobe while longer range patterns indicate marked compression of this lobe, to widths as small as one half the standard lobe width. Comparison with meteorological soundings is given and the relationship of the measurements to anomalous propagation conditions close to the water surface is indicated.

### 23. VHF TROPOSPHERIC RECORDING MEASUREMENTS OF PLANE AND CIRCULAR POLARIZED WAVES IN THE GREAT LAKES AREA

J. S. Hill
(United Broadcasting Company, Cleveland, Ohio)
and
G. V. Waldo
(Federal Communications Commission, Washington, D. C.)

Results are given of measurements of characteristics of circularly polarized vhf waves propagated through the troposphere. Continuous field intensity recordings of horizontal and vertical components as well as the circularly polarized field over a 125-mile path from Columbus, Ohio, are given. Results are compared with recordings of plane-polarized wave propagation over the same path and over a different path.

### 24. THE EXPERIMENTAL AND THEORETICAL STUDY OF IONOSPHERIC ABSORPTION AT 150 KC

A. H. Benner
(RCA Victor Division, Camden, N. J.)

Analytical expressions for vertical incidence ionospheric absorption in $E$ layer are obtained from a study of the complex index of refraction for 150 kc. The double parabolic approximation to the Chapman electron distribution and an exponential frequency variation with height are used. Calculations of the total absorption made by summing the contributions of the $E$ region and the nondeviating $D$ region indicate that the values of $E$-region absorption depend greatly on the characteristics of this region near the maximum of ionization. It is tentatively concluded that the $D$ region is not a pure electron layer and that it probably does not exhibit a distinct ionization maximum. Typical vertical incidence 150-kc absorption data are presented and the results of analysis of 17 months' data are discussed.

### 25. THE MEASUREMENT OF THE POLARIZATION OF IONOSPHERIC REFLECTIONS AT LOW FREQUENCIES

R. A. Helliwell, A. J. Mallinckrodt, D. A. Campbell, and W. Snyder
(Stanford University, Stanford, Calif.)

A relatively simple method for measuring the polarization of pulse echoes is described not involving the use of frequency converters or phase shifters. Examples are given using typical $E$-layer echoes at 100 and 340 kc. Vertical incidence results indicate both directions of rotation of the polarization ellipse at 310 to 340 kc but only left-handed rotation at 100 kc. At times marked changes in polarization occurred within a few rf cycles. Cases of echo splitting in which the component polarizations are opposite in sense of rotation are considered in terms of ray theory.

### 26. THEORETICAL AND EXPERIMENTAL INVESTIGATION OF THE POLARIZATION OF LONG WAVES REFLECTED FROM THE IONOSPHERE

J. M. Kelso and H. J. Nearhoof
(The Pennsylvania State College, State College, Pa.)

Methods are given for determining the relations between measurable quantities of down-coming echo polarization and a complex polarization $R$ defined by Appleton and Hartree. Calculations are made for 150 kc at State College, Pa. A cross-loop polarimeter is described. Polarization results indicate a definite diurnal trend with limiting polarization taking place for collisional frequency values greater than the critical.

### 27. POLARIZATION MEASUREMENTS OF LOW-FREQUENCY ECHOES

E. L. Kilpatrick
(National Bureau of Standards, Washington, D. C.)

A method for determination of polarization in a vertically incident down-coming radio wave is described using crossed dipoles and pulses. A gating arrangement permits selection of 25 microsecond samples for any desired portion of the echo, thus permitting independent observations of ordinary and extraordinary components. Experimental results indicate the polarization to be in general elliptical, with left-handed sense of rotation. The rapid time variation of polarization which sometimes occurs within a single echo suggests the presence of a composite reflection unresolved in time.

### 28. A METHOD FOR OBTAINING THE WAVE SOLUTIONS OF IONOSPHERICALLY REFLECTED LONG RADIO WAVES INCLUDING ALL VARIABLES AND THEIR HEIGHT VARIATION

J. J. Gibbons and R. J. Nertney
(The Pennsylvania State College, State College, Pa.)

A method is presented for obtaining approximate solutions to the wave equation in the form of a linearly independent pair involving a parameter $p$ which is a complex function of the height. The solutions reduce to WKB solutions for regions for small partial reflection but differ considerably in region of considerable reflection. Examples are given for various configurations including derivation of reflection coefficients. Advantages include the simplicity of application and no requirement for explicit knowledge of the refractive index off the axis of reals.

### 29. ON THE THEORY OF ANTENNA BEAM-SHAPING

A. S. Dunbar
(Stanford Research Institute, Stanford, Calif.)

The diffraction pattern of an aperture with specified amplitude distribution and phase aberration is examined for two cases. The effects of these parameters on the beam shape and its control are discussed. A general formulation of Chu's method[1] for an amplitude distribution on a curved surface is developed. Theoretical results are applied to the design of progressive-phase antennas for both control variation of the amplitude distribution and the controlled variation of phase. Experimental results obtained with long-slot antennas, corrugated-surface antennas, slot-array antennas, and a dielectric slab on a curved surface, are presented.

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[1] L. J. Chu, "Microwave Beam-Shaping Antennas," Tech. Rpt. No. 40. Research Lab. of Electronics, MIT, Cambridge, Mass.; 1947.

### 30. BEAM SHAPING IN DOUBLY CURVED REFLECTOR SYSTEMS EMPLOYING QUASI-POINT SOURCES

A. E. Marston

(Naval Research Laboratory,
Washington, D. C.)

Techniques for designing doubly curved reflectors for shaped beams, developed by Chu and Silver, are extended to embrace the case of the so-called quasi-point source. Such sources in general are identified with a one-parameter family of normal congruences which includes the spherical and cylindrical congruences as special cases. An explicit design procedure is given and experimental confirmation presented.

### 31. LACK OF UNIQUENESS IN ANTENNA PATTERN SYNTHESIS METHODS AND THE RELATED ENERGY STORAGE CONSIDERATIONS

T. T. Taylor

(Hughes Aircraft Company,
Culver City, Calif.)

A modification of a nonsynthesis method for aperture-type antennas is discussed and it is shown how, within directivity limitations, high-stored energy solutions can be avoided. The synthesis problem becomes more difficult if the objective is to find an excitation function whose Fourier transform shall approximate a given function with a specified interval. This problem can be solved by methods analogous to those discussed by Whinnery and the author in regard to arrays of discrete elements. Extent of the lack of uniqueness of the solution is discussed.

### 32. FEED PROBLEMS IN BROADBAND ANTENNA ARRAYS

W. R. LePage and R. F. Gates

(Syracuse University, Syracuse, N. Y.)

The input admittance of an element of the array is analyzed to show the effect of coupling to adjacent elements. This is particularly important when adjacent elements are fed in different phase. The analysis is carried out by neglecting coupling to elements farther from a given element than its immediate neighbors. An equivalent circuit is evolved and analyzed for the input admittance to such an element. Also, the range to be expected for the standing-wave ratio on a line feeding a tapered section is discussed in terms of the impedance transformation of the section, and the standing-wave ratios on the various lines which it feeds. Among other things, it is shown that the standing-wave ratio on the feed to the tapered section is always less than the average of the standing-wave ratios on the lines leading from it, if the tapered section has the proper impedance ratio.

### 33. SECOND-ORDER BEAMS OF SLOTTED WAVEGUIDE ARRAYS

H. Gruenberg

(National Research Council,
Ottawa, Ont., Canada)

An array of longitudinal shunt slot in the broad face of a rectangular waveguide is studied. It is shown that the radiation pattern of such an array will contain small second-order beams even for slot spacings less than a free-space wavelength. These beams are usually split with a minimum or even a null in the plane containing the axis of the array and the maximum of the main beam. Second-order beams may be as large as 10 per cent of the main beam (in voltage) for typical arrays about 50 wavelengths long, and even larger for shorter arrays. Methods of suppressing these beams are discussed and experimental confirmation is given.

### 34. VERY-HIGH-FREQUENCY PROPAGATION IN THE EQUATORIAL REGION

O. P. Ferrell

(Radio Magazines, Inc.,
Philadelphia, Pa.)

Radio amateur observations in the frequency range 50.0–54.0 mc are being collected throughout the Western Hemisphere. Sky-wave propagation in the geographic tropical zone may be divided into three categories, which are: (1) sporadic-$E$ reflections; (2) $F_2$-layer reflections, particularly associated with ionospheric storms in the temperate latitudes; (3) post-sunset $F_2$-region scatter. The effect of these modes upon vhf propagation are summarized with particular emphasis placed upon the hitherto unreported post-sunset scatter. The results of the present study indicate that the scattering is observed over fairly long paths, but is restricted to those months around the equinoxes

### 35. SOUTHERN EXTENT OF AURORA BOREALIS IN NORTH AMERICA

C. W. Gartlein and R. K. Moore

(Cornell University, Ithaca, N. Y.)

Results are presented for the first eleven years of a study of the frequency of overhead auroras in North America as a function of latitude in a region south of the auroral zone. The data have been averaged in various ways so that monthly and annual variations are demonstrated. It appears that there is a relatively constant level of auroral activity throughout the year in the region 58°–60° geomagnetic latitude, while auroras appearing overhead south of these latitudes are more frequent during equinoctial periods. Auroras have been seen as far south as 52° during this period every month of the year. Correlation of auroral frequency with sunspot number is not high on a month-by-month or three-month running mean basis.

### 36. A VHF PROPAGATION PHENOMENON ASSOCIATED WITH THE AURORA

R. K. Moore

(Cornell University, Ithaca, N. Y.)

Anomalous propagation observed by radio amateurs at 28 to 148 mc during displays of Aurora Borealis is described. This phenomenon is characterized by the following features: (1) a very high fading rate, such that voice modulation is rendered unintelligible; (2) directional antennas give best results when pointed north (toward the aurora); (3) lack of skip effect; (4) little change in polarization. Results indicate this effect to be most common with auroral displays extending below 56° geomagnetic latitude.

### 37. PHASE VELOCITY OF VERTICALLY POLARIZED ELECTROMAGNETIC WAVES IN THE DIFFRACTION REGION AT THE SURFACE OF A SPHERE

Henry Lisman

(Signal Corps Engineering Laboratories,
Fort Monmouth, N. J.)

Based on the work of Eckersley and Millington, an investigation was carried out to determine phase velocity of electromagnetic waves over the surface of a sphere. Numerical calculations were carried on for 5.305 mc on various assumptions as to values of dielectric constant and conductivity. The velocity was found insensitive to changes in ground constants. Changes in assumed radius of curvature were found to affect the velocity, thereby indicating a substantial difference over a region where the curvature varies considerably to that over a smooth earth.

### 38. MAGNETO-IONIC MULTIPLE SPLITTING DETERMINED WITH THE METHOD OF PHASE INTEGRATION

Wolfgang Pfister

(Air Force Cambridge Research
Laboratories, Cambridge, Mass.)

The method used is a first-order WKB approximation with an integration path in the complex height plane. Suitable paths of integration in the Riemann surface determine the reflection coefficient for the various magneto-ionic components. Numerical computations have been carried out for a Chapman distribution of electrons and an exponential decrease of collisional frequency located in a moderate magnetic latitude. The results are represented in a sweep-frequency picture of virtual height and absorption for five fundamental modes.

### 39. ELECTROMAGNETIC ENERGY DENSITY AND FLUX

C. O. Hines

(Canadian Defence Research Board,
Ottawa, Ont., Canada)

J. C. Scott and H. G. Booker by different methods have obtained discrete formulas for energy flow in the ionosphere when absorption is considered. Investigation of the bases of the two methods revealed that the usual expressions for energy density and flux lead to difficulties which are avoided by alternative expressions based on an equally valid conservation theorem. These alternative formulas are shown to lead to much neater results, not only for the dissipative case, but for all problems in plane wave propagation, and also give the group speed for the speed of the energy.

### 40. ASYMPTOTIC SOLUTION OF MAXWELL'S EQUATION

Morris Kline

(New York University, New York, N. Y.)

A method is given which improves upon geometrical optics solutions of electromagnetic problems by establishing asymptotic series for $E$ and $H$. The successive coefficients of the two series are obtained by solving two recursive systems of first-order ordinary differential equations. These differential equations must be solved along each system of rays. At any point in space each coefficient of the series is the sum of the solutions of the differential equations corresponding to rays passing through the point.

### 41. CURRENT DISTRIBUTIONS ON LARGE REFLECTING CYLINDERS

H. J. Riblet

(Microwave Development Laboratories, Inc., Waltham, Mass.)

This preliminary report describes calculations and measurements made with the intention of determining the current distribution on an infinite circular cylinder, whose radius is $12\lambda 2\pi$, when placed in a plane wave. Two cases are considered. For Case I, the incident electric vector is polarized parallel to the axis of the cylinder, while for Case II, the incident magnetic vector is polarized in that direction. Six different current distributions are compared in amplitude and phase for both cases and some general conclusions are reached.

### 42. APERTURE PHASE ERRORS IN MICROWAVE OPTICS

K. S. Kelleher

(Naval Research Laboratory, Washington, D. C.)

The fundamental difference between optics and microwave optics lies in the fact that the former stresses the aberrations resulting from imperfect focussing, while the latter is more concerned with aberrations in terms of the deviation of a wavefront from a plane. An indication of the relation between the phase errors which describe this deviation and the radiation pattern is made by considering the case of a paraboloidal reflector with feed displaced from the focal point and the case of a spherical cap reflector with feed on axis. In these instances, consideration is given to the amplitude of intensity across the aperture since phase errors are relatively unimportant when they occur in the region of the aperture in which the intensity is low.

### 43. REVIEW OF SPHERICAL REFLECTOR RESEARCH AT AFCRL

J. E. Walsh

(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

Spherical reflectors have been investigated because of their superiority to the paraboloid in wide-angle scanning potentialities, even though they are subject to spherical aberration. The optimum location of a point source feed was determined and experimental measurements made on patterns. Line source feeds with nonconstant phase gradients were also investigated, resulting in an aberrationless optical system with a favorable $f/d$ ratio which can scan a solid angle of at least 60°. Successful experimental tests of these line sources have been made.

### 44. REVIEW OF RECENT METAL LENS RESEARCH AT AFCRL

John Ruze

(Air Force Cambridge Research Laboratories, Cambridge, Mass.)

Lenses of the "constrained" type have been investigated in which the ray direction is determined by the metal-plate configuration. Lenses of this type lend themselves readily to wide-angle analysis, and can be designed to focus perfectly at two symmetrically placed off-axis points. These corrected lenses have good scanning properties and have been successfully constructed for use with both line and point sources. Work has also been done on achromatic lenses in which electromagnetic energy in the *TEM* mode of transmission is guided through the plates. Two methods of focussing by adjusting the path length through each pair of plates are described.

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## ACOUSTICS AND AUDIO FREQUENCIES

**016:534** **1042**

**References to Contemporary Papers on Acoustics**—A. Taber Jones. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 888–894; November, 1950.) Continuation of 1 of February.

**534:061.4** **1043**

**The Audio Fair Review**—(*Audio Eng.*, vol. 34, pp. 24–30, 54; December, 1950). Brief descriptions of loudspeakers, amplifiers, pick-ups, recorders, and so forth, exhibited at the Second Audio Fair.

**534.21:551.463** **1044**

**The Dependence of Shadow-Zone Sound on the Surface-Sound Velocity Gradient**—R. W. Morse. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 857–860; November, 1950.) Sound penetrating the so-called "shadow zone" of ocean acoustics is investigated by ray-theory methods. On the assumption that the sound-velocity gradient approaches zero in a surface layer a few feet deep, sound levels in the shadow are derived of the same order as those measured at a frequency of 24 kc.

**534.313.4:534.373** **1045**

**Boundary-Layer Attenuation of Higher-Order Modes in Rectangular and Circular Tubes**—R. E. Beatty, Jr. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 850–854; November, 1950.) The boundary-layer attenuation caused by viscous and thermal losses of all higher-order modes of propagation in rigid-walled tubes is calculated, using the concept of boundary-layer admittance introduced by Cremer. The symmetrical (0, *n*) circular-tube modes are the least attenuated of all higher-order modes in both rectangular and circular tubes; above a certain frequency the attenuation is less than that for a plane wave.

**The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.**

**534.23/.24** **1046**

**The Equivalent Circuit for the Transmission of Plane Elastic Waves through a Plate at Oblique Incidence**—W. T. Thomson. (*Jour. Appl. Phys.*, vol. 21, pp. 1215–1217; December, 1950.) Formulas are derived for the transmission and reflection coefficients of a plate with different liquids in contact with its two faces. A simple equivalent circuit involving impedances which are functions of the angle of incidence is given for a plate transmitting both the dilatation and the shear wave.

**534.26** **1047**

**On the Diffraction of a Plane Sound Wave by a Paraboloid of Revolution**—C. W. Horton and F. C. Karal, Jr. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 855–856; November, 1950.) "An infinite set of equations is obtained for the coefficients of the pressure fields inside and outside the paraboloid. For the particular case of a rigid paraboloid the coefficients of the series for the scattered field can be obtained explicitly. The ratio of the total pressure at the nose of the paraboloid to the pressure of the incident wave is shown as a function of the shape of the paraboloid for a plane wave whose wave front is perpendicular to the axis of revolution. The same quantity is plotted for the case of a sphere whose radius of curvature is equal to that of the paraboloid at the nose."

**534.321.9:534.22:577** **1048**

**The Velocity of Sound through Tissues and the Acoustic Impedance of Tissues**—G. D. Ludwig. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 862–866; November, 1950.)

**534.321.9:534.373.2** **1049**

**Effects of Humidity on Ultrasonic Absorption in Air at 1.46**—G. S. Verma. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 861–862; November, 1950.) Using the acoustic resonator interferometer developed by Hubbard (1932 Abstracts, pp. 171 and 647), the absorption coefficient ($\alpha$) was determined for air of relative humidity ranging from zero to 84 per cent. At the frequency used, a maximum value $\alpha 0.52$ cm$^{-1}$ was found for a relative humidity of 46 per cent. This confirms results obtained at lower frequencies by Knudsen, Pielemeier, and Richardson.

**534.321.9:577** **1050**

**Physical Factors Involved in Ultrasonically Induced Changes in Living Systems: Part 1. Identification of Non Temperature Effects**—W. J. Fry, V. J. Wulff, D. Tucker, and F. J Fry. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 867–876; November, 1950.) Changes in physical variables accompanying high-intensity ultrasonic disturbances in liquid media are discussed. The significance of temperature changes is investigated experimentally for nerve and other tissues of frogs and crayfish, at frequencies of 975 to 980 kc; factors other than absorption of acoustic energy, that is, cavitation, are found to be operative. The method involves determination of the acoustic absorption coefficient of the specimens from measurements of initial rate of change of temperature.

**534.756** **1051**

**Theory of Operation of the Cochlea: a contribution to the Hydrodynamics of the Cochlea**—O. F. Ranke. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 772–777; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950.

**534.756** **1052**

**Theory of the Acoustical Action of the Cochlea**—J. Zwislocki. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 778–784; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950.

**534.771** **1053**

**The Electrical and Acoustical Performance of some Commercial Audiometers**—K. C. Morrical, R. W. Benson, and H. Davis. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 843–847; November, 1950.)

**534.771.089.6** **1054**

**Calibration of Audiometers**—E. L. R. Corliss and W. F. Snyder. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 837–842; November, 1950.) An account is given of the technique developed at the National Bureau of Standards. Standard sources are described, and the physical basis of the method is discussed. Limitations on the validity of the threshold data are pointed out. Location of causes of defective operation in audiometers is facilitated by reverence to earphone response data.

**534.78:519.272.119** **1055**

**Correlation Function Analysis**—Kraft. (*See* 1179.)

**534.78:519.272.119** **1056**

**Autocorrelation Analysis of Speech Sounds**—K. N. Stevens. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 769–771; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. The short-time autocorrelation function is defined and is used to analyze various speech sounds. Equipment used at the Massachusetts Institute of Technology is described.

**534.76:621.314.26** **1057**

**An Apparatus for Speech Compression and Expansion and for Replaying Visible Speech Records**—F. Vilbig. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 754–761; November, 1950.) A paper presented at the Speech Communication

Conference at Massachusetts Institute of Technology, 1950. *See* also 2693 of 1950.

534.782 1058

**An Electrical Vocal System**—L. O. Schott. (*Bell Lab. Rec.*, vol. 28, pp. 549–554; December, 1950.) An outline description of apparatus arranged to form an electrical analogue of the human vocal system. The "vocal tract" comprises an artificial transmission line of 24 L sections. The "tongue hump" is an auxiliary inductor of adjustable value ("hump-height") which can be switched in series with any chosen inductive series element of the transmission line. The "lip opening" is an adjustable inductor which terminates the line and across which the loudspeaker is connected. Separate generators are provided for producing vowels and unvoiced fricatives, while further arrangements enable pitch and inflection to be varied and other vocal features to be reproduced.

534.784 1059

**System-Function Analysis of Speech Sounds**—W. H. Huggins. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 765–767; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. "The analysis of speech sound is facilitated if the sound is considered to be the response of a slow-time-varying linear system to appropriate excitations. The linear system is characterized by one or more system functions which may be represented most naturally as the sum of complex exponential terms having complex frequencies corresponding to the various formants. Thus, a generalized frequency analysis is made not of the speech wave itself but rather of the system function that shaped that wave. On the other hand, the excitation is best analyzed by autocorrelation methods."

534.861:534.76 1060

**Stereophonic Broadcasting of Music**—J. Bernhart and J. W. Garrett. (*Toute la Radio*, no. 150, pp. 353–358; November, 1950.) Discussion of methods of obtaining "relief" effect in sound reproduction, including an analysis of directional discrimination in audition and a description of the principle of "directed stereophony." Variation of intensity itself is sufficient to convey a sense of spatial movement of the sound source, particularly for low frequencies. This can be accomplished using a single microphone and two separate modulation channels the levels of which are adjusted independently: a 5-db alteration between the two levels corresponds to a radial movement of 30° angular displacement being a linear function of the intensity difference. Phase variation has a similar effect but necessitates a more complex system in practice. Displacement in depth is conveyed by use of an echo room. The effects of combination and rapidity of these adjustments are discussed with reference to an experimental broadcast transmission from Paris [*see* 1061 (Aisberg)].

534.861:534.76 1061

**Improved Stereophony**—E. Aisberg. (*Wireless World.* vol. 56, pp. 327–330; September, 1950.) An account of experiments in France, using simultaneous transmissions from the Parisian and Paris Inter broadcasting stations. For reception two loudspeakers spaced 5 to 7 feet apart were required, the listeners being 7 to 10 feet away. The two receivers receive signals from separate channels and by suitably controlling the modulation depth for the two channels, a realistic impression of movement of the original source of sound is given to the listeners. Modulation of the two channels was recorded on double-track magnetic tape. *See* also 1060 above.

621.314.2.029.3:621.3.018.78† 1062

**The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers**—Macfayden. (*See* 1087.)

621.392.52 1063

**Unusual Ladder Filter. Applications in Audio and Radio Circuits**—Davey. (*See* 1112.)

621.395.623.45 1064

**An Ultrasonic Underwater 'Point Source' Probe**—L. Fein. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 876–877; November, 1950.) An ultrasonic receiver uses a conical piezoelectric probe made of ADP. Its sensitivity over the frequency range 1.235 to 1.280 mc was determined by means of a reciprocity calibration, an average value of 0.012 $\mu v$ per dyne per cm$^2$ being obtained. The measured directional pattern does not very with position in the field, indicating that the probe is effectively a point receiver.

621.395.625.3/.6(41) 1065

**Magnetic-Sound-Film Developments in Great Britain**—O. K. Kolb. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 55, pp. 496–508; November, 1950.) The introduction of this type of recording and reproducing apparatus in Great Britain is reviewed. Types of magnetic film available and their characteristics are described. The general circuit arrangements and details of apparatus are given together with a description of special apparatus that has been used for adding a visible record to the invisible magnetic sound track. Jointing processes and a method for bulk erasure of records on magnetic-film stock are discussed.

621.396.645.029.3:621.385.3/.4 1066

**A Comparison of Triodee and Beam-Tetrodes as Power Output Valves in Audio Amplifiers**—Bruckmann, Carey, and Fuller. (*See* 1123.)

## ANTENNAS AND TRANSMISSION LINES

621.392 1067

**Phase Distortion in Feeders**—L. Lewin, J. J. Muller, and R. Basard. (*Elect. Commun.* (London), vol. 27, pp. 320–323; December, 1950.) Reprint. *See* 2125 of 1950.

621.392:621.3.012.1/.2 1068

**Geometrical Solution of some Transmission-Line Problems**—Q. C. Gupta and F. Duerden. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 525–529; December, 1950.) A description of the vector-diagram approach, which is said to provide a better physical picture than analysis based on differential equations. The method is illustrated by an analysis of stub matching and a verification of the principles of the circle-diagram calculator.

621.392:[621.3.015.71:621.314.2 1069

**Pulse Transients in Exponential Transmission Lines**—In 280 of March, for "E. R. Schatz" please read "E. R. Schatz and E. M. Williams."

621.392:621.3.09 1070

**The Propagation of Transient and Periodic Waves along Transmission Lines**—R. Pélissier. (*Rev. gén. Elect.*, vol. 59, pp. 379–399, 437–454 and 502–512; September to November, 1950.) A general theory of the mode of propagation and of the different causes of attenuation of electric waves is developed, using the symbolic calculus. The theory generalizes those of Carson and Pollaczek. Approximate formulas are derived and the limits of their applicability are discussed. Applications of the theory to cases of practical importance include, (*a*) calculation of charge and current distribution along an infinitely long line after application of a step voltage, (*b*) corresponding calculations for a line of finite length, including discussion of reflection at terminations of various types, (*c*) reflection at line discontinuities, (*d*) discussion of causes of wave attenuation, (*e*) propagation of high-frequency currents and of surges due to lightning. The theory shows that the explanation of the attenuation of such surges depending on corona effects is inexact. A new explanation is developed and confirmed by experimental results. *See* also 2972 of 1950.

621.392.011.2:621.3.012.3 1071

**Transmission-Line Impedance Curves**—H. A. Wheeler. (Proc. I.R.E., vol. 38, pp. 1400–1403; December, 1950.) A universal family of curves gives the wave resistance, inductance, and capacitance of transmission lines with 1, 2, or 4 wires. Each curve represents one field pattern. An expanded scale is given for the region of very low impedance. The seven field patterns are for the cases of conductors of circular cross section in or near shields such as a plane, inclined or parallel planes, a trough, or a coaxial conductor.

621.392.4 1072

**A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped *LC* Circuits**—van Weel. (*See* 1107).

621.396.67 1073

**Theory of Collinear Antennas**—R. King. (*Jour. Appl. Phys.*, vol. 21, pp. 1232–1251; December, 1950.) The analysis of a system of collinear antennas is practicable only if the transmission lines feeding the system are in the neutral plane of the electromagnetic field. Two cases are considered: a single center-driven unit, (*a*) with two symmetrical collinear parasitic elements, and (*b*) with the outer elements driven from the center unit by means of phase-reversing stubs. Approximate expressions are obtained for the currents and for the mutual and self impedances of the elements. The particular case of $\lambda/2$ dipoles is treated in detail. The advantages of the collinear array with phase-reversing stubs as compared with arrays in which each element is center driven are discussed.

621.396.67 1074

**Elliptically Polarized Waves**—L. Hatkin. (Proc. I.R.E., vol. 38, p. 1455; December, 1950.) Comment on 3343 of 1948 (Sichak and Milazzo).

621.396.67 1075

**On the Radiation Patterns of Dielectric Rods of Circular Cross Section—the $\text{TM}_{0,1}$ Mode**—C. W. Horton, F. C. Karal, Jr., and C. M. McKinney. (*Jour. Appl. Phys.*, vol. 21, pp. 1297–1283; December, 1950.) Radiation patterns were measured for five rods of diameter *d* equals $0.87\lambda_0$ and lengths $2\lambda_0$ to $10\lambda_0$. The theoretical radiation patterns were computed by means of equivalent-surface electric and magnetic currents. Excellent agreement is obtained when it is assumed that the diameter of the surface on which these currents are distributed is 0.65*d*. Representative experimental and theoretical patterns are shown for the full 360°.

621.396.67:621.397.6 1076

**Temporary Vision Aerial**—F. D. Bolt. (*Wireless World*, vol. 57, pp. 13–14; January, 1951.) A brief description of a temporary folded dipole used for radiating the vision signal from the London television transmitter (frequency, 45 mc; mean power, 7.5 kw) while the permanent system was being overhauled. Two coaxial feeders, about 400 feet long, one being a half wavelength longer than the other, were used as a "binocular pair" to energize the antenna. A central element, of *Al* tubing, surrounded by three driven elenents, gave a driving-point impedance of $168\Omega \pm 10$ per cent over the frequency band 42 to 48 mc. Test transmissions produced pictures nearly as good as those from the original antenna.

621.396.67.029.64 1077

**Microwave-Aerial Radiation Patterns**—D. G. Kiely and A. E. Collins. (*Wireless Eng.* vol. 28, no. 328, pp. 23–29; January, 1951.) The errors involved in measuring the side-lobe amplitudes of 3-cm fan-beam antennas are discussed. Site errors and instrumental errors are considered in detail. Experimental evidence shows that the largest errors are caused by reradiation from trees, hedges, and similar obstacles near the measurement sites. For two typical field sites, when three slightly different frequencies were used, the standard deviation of site errors in measurements of side lobes in the radiation patterns of cheese antennas was 1.4 db. The design of antennas to conform to specifications for slide-lobe magnitude must make allowance for such errors.

621.396.677 1078
**An Achromatic Microwave Antenna**—N. I. Korman and J. R. Ford. (Proc. I.R.E., vol. 38, pp. 1455–1456; December, 1950.) Comment on 1345 of 1950 (Ruze).

621.396.677:538.56 1079
**Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates**—Berz. (*See* 1138.)

621.396.677:621.396.932.1 1080
**Cheese Aerials for Marine Navigational Radar**—D. G. Kiely, A. E. Collins, and G. S. Evans. (*Proc. IEE* (London), Part III, vol. 98, pp. 37–45; January, 1951.) A consideration of the problem of the suppression of side lobes in the horizontal radiation pattern of fan-beam antennas. The theory of the primary feed requirements of cheese and half-cheese antennas is outlined and the effect of the secondary radiator on the radiation of the primary feed, and hence on the distribution of energy across the aperture, is analyzed. The side lobes of a symmetrical cheese antenna cannot be reduced much more than 23 db below the main-beam level. The design and performance of a half-cheese antenna having side lobes 30 db below the main-beam level are described.

**Book**

621.396.67 1081
**Antenna Theory and Design, Vols. 1 and 2 [Book Review]**—H. P. Williams. Publishers: Pitman and Sons, London, 1950, 142 pp. and 522 pp., 21s. and 63s. (*Jour. Brit. IRE*, vol. 10, p. vii; December, 1950.) The first volume, which is independent of the second, deals with the theory of antenna. The treatment is necessarily mathematical, but numerous diagrams are used in illustration of the theory. The second volume is concerned primarily with antenna design. All types are considered and many curves, diagrams, and photographs are given. A full bibliography is included in both volumes.

## CIRCUIT AND CIRCUIT ELEMENTS

537.312.6:621.315.592† 1082
**Thermistors**—Please note that the above UDC number is now used for thermistors in place of 621.316.86 used hitherto.

537.312.6:621.315.592† 1083
**A Relationship between Resistance and Temperature of Thermistors**—G. Bosson, F. Gutmann, and L. M. Simmons. (*Jour. Appl. Phys.*, vol. 21, pp. 1267–1268; December, 1950.) The equation $\log R = A + B/(T + \Theta)$ for the resistance $R$ of a thermistor at $T$°K is proposed. Least-square analyses of the most precise resistance temperature data available for three different thermistor materials show that the equation is a considerable improvement over its predecessors; the standard relative errors of fit are 0.4 per cent, 0.17 per cent, and 0.91 per cent.

621.3.011.3/.4:621.3.012.3 1084
**Reactance Chart**—H. A. Wheeler. (Proc. I.R.E., vol. 38, pp. 1392–1397; December, 1950.) An extension of the range of the normal chart involving reactance, frequency, inductance, and capacitance, with added scales for susceptance, wavelength, and time constant. Simple geometrical patterns are given which enable the chart to be used for the direct solution of problems such as the bandwidth of a resonant circuit, wave impedance, and delay of a transmission line, and design of constant-*k* filter half sections.

621.3.015.7†:621.3.087 1085
**A Stable Ninety-Nine Channel Pulse-Amplitude Analyser for Slow Counting**—D. H. Wilkinson. (*Proc. Camb. Phil. Soc.*, vol. 46, Part 3, pp. 508–518, July, 1950.) The input pulse is converted into a series of pulses of standard height, their number being proportional to the amplitude of the input pulse. Registration is then done by passing the series of pulses into two decade counting rings in cascade; the first ring passing a pulse to the second ring each time it has itself made a complete circuit. Attached to each element of the rings is a telephone message register which operates only when the pulse counting stops at the associated element. The analyzer accepts pulses at a maximum rate of 10 per second. Its calibration is linear to within 1 per cent, and independent of pulse shape, provided the pulse rise time is between 1 and 50 $\mu$s.

621.3.016.352† 1086
**Relation of Nyquist Diagram to Pole-Zero Plots in the Complex Frequency Plane**—W. W. Harman. (Proc. I.R.E., vol. 38, pp. 1454–1455; December, 1950.) Two criteria for the stability of a closed-loop system consisting of an amplifier of gain $A$ with fractional feedback $\beta$ are: (*a*) the complex gain $A$ to $(1\text{-}A\beta)$ must not have a pole in the right half of the complex frequency plane; (*b*) the Nyquist polar plot of the loop gain $A\beta$ must not enclose the point (1, 0). The equivalence of these conditions is demonstrated by reference to a 4-stage *RC* amplifier.

621.314.2.029.3:621.3.018.78† 1087
**The Calculation of Waveform Distortion in Iron-Cored Audio-Frequency Transformers**—K. A. Macfadyen. (*Proc. IEE* (London), Part II, vol. 97, pp. 809–811; December, 1950.) "In the design of transformers, the prediction of the amount of waveform distortion from a knowledge of the properties of the core mterial becomes complicated if the preceding circuit has a high impedance. A theorem is given showing that, if suitable distortion and other measurements are made on a sample core of the material to be used and the distortion factor plotted as a function of two suitably chosen variables, then this information is sufficient to enable the distortion arising from any transformer in any resistive circuit to be calculated exactly."

621.314.22:621.395.641.1 1088
**Double-Humped Resonance Curves of Repeaters**—G. Schmitt and H. Schrag. (*Fernmeldetech. Z.*, vol. 3, pp. 422–427; November, 1950.) Experiment shows that the two resonance peaks in the output from the divided secondary winding of a repeater are due to the different winding capacitances of the two windings $w_2$ and $w_3$. One peak, is eliminated by shunting $w_2$ with a capacitor, connecting the outer end of $w_2$ to the inner end of $w_3$ and earthing the outer end of $w_3$. Theory of the method is outlined.

621.314.224 1089
**Dielectric Admittances in Current Transformers**—A. H. M. Arnold. (*Proc. IEE* (London), Part II, vol. 97, pp. 727–732. Discussion, pp. 733–734; December, 1950.) For a current transformer with appreciable capacitance between windings, five ratios may be defined without knowledge of the external circuits, but only two of these ratios can be measured directly with a simple bridge. Formulas are developed for evaluation of all the ratios from simple bridge measurements, and the formulas are tested experimentally. A simple test is described for determining quickly whether the effect of capacitance is appreciable in a transformer of unknown characteristics, and methods of construction of low-capacitance transformers are described.

621.314.224 1090
**The Effect of Capacitance on the Design of Toroidal Current-Transformers**—A. H. M. Arnold. (*Proc. IEE* (London), Part II, vol. 97, pp. 797–809; December, 1950.) For high-accuracy operation at currents below 5 amps or frequencies above 50 cs, transformer performance is dependent not only on the magnitude of the self-capacitances of the windings and their mutual capacitance, but also on the distribution of the capacitance between windings. The normal method of layer winding gives an unsatisfactory concentrated distribution, but a form of bank winding is described which gives a symmetrical distributed capacitance between windings, resulting in improved performance. If the thickness of the insulation between windings is increased to about one eighth of the mean depth of the secondary winding and an insulant of low permittivity is used, further improvement in performance is obtained. The performance of multi-ratio transformers is considered and good winding arrangements are suggested.

621.314.3† 1091
**Feedback in Magnetic Amplifiers**—A. S. Fitzgerald. (*Elec. Commun.* (London), vol. 27, pp. 298–319; December, 1950.) Reprint. *See* 2731 of 1949.

621.314.3† 1092
**Magnetic-Amplifier Characteristics**—L. A. Finzi and D. C. Beaumariage. (*Elec. Eng.* vol. 69, pp. 1109–1115; December, 1950.) Essential text of American Institute of Electrical Engineers 1950 Summer and Pacific General Meeting paper. Dimensionless curves are presented for the determination of the steady-state output characteristics of magnetic amplifiers in terms of design parameters. Correction factors for the feedback ratio are calculated. Results obtained by this method agree well with experimental results for amplifiers with cores of gradually varying incremental permeability.

621.314.3† 1093
**A Theoretical and Experimental Study of the Series-Connected Magnetic Amplifier**—H. M. Gale and P. D. Atkinson. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 302–303; November, 1950.) Discussion of paper abstracted in 2729 of 1949.

621.314.3† 1094
**Magnetic Amplifiers**—A. G. Milnes. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 302–303; November, 1950.) Discussion of paper abstracted in 2728 of 1949.

621.314.3† 1095
**A Theory of the Series Transductor**—C. S. Hudson. (*Proc. IEE* (London), Part II, vol. 97, pp. 751–755; December, 1950.) "The theory of the series transductor with natural excitation is discussed and general equations for load and control current are derived. The effect of phase shift between the supply voltage and that across the transductor is taken into account, and the transient changes which occur at the moment of transductor cut off are explained. The effect on the transductor characteristic of the shape of the $B/H$ curve for the core material is considered, and the analysis is extended to enable an experimental $B/H$ curve to be used in the derivation of the transductor characteristic. It is shown that, with 100 per cent self excitation, the transductor characteristic approximates in shape to the upper part of the $B/H$ curve."

621.318.4 1096
**Design of Iron-Cored Inductances carrying D. C.**—N. H. Crowhurst. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 516–523; December, 1950.) A series of charts based on

the properties of typical transformer iron permit the rapid design of inductors for many purposes. From the dc polarizing current and the required inductance and resistance, the optimum core size, number of turns and width of air gap may be readily determined. Data are tabulated for laminations of stock sizes.

621.318.423 1097

**A 100-Kilowatt Water-Cooled Solenoid**—J. M. Daniels. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1028–1034; 1st December, 1950.) Design data and construction details for an air-cored solenoid, producing a field of 14.7 kilogauss, uniform to within 0.5 per cent in a cylindrical volume of length 6 cm and diameter 4 cm.

621.318.423.011.3:621.3.012.3 1098

**Inductance Chart for Solenoid Coil**—H. A. Wheeler. (Proc. I.R.E., vol. 38, pp. 1398–1400; December, 1950.) A simple chart relates inductance, coil over-all dimensions, and winding density. Logarithmic scales covering several decades suffice for almost all practical applications.

621.318.423.015.33 1099

**The Natural Frequencies of Single-Layer Solenoids subjected to Voltage Surges**—B. Heller, J. Hlávka, and A. Veverka. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 24, pp. 951–957; 26th November, 1949. In German.) Formulas for the natural frequencies are developed, taking into account the mutual inductance between turns, for the two cases where the far end of the coil is (*a*) earthed or (*b*) free. In case (*a*) the spatial distribution of voltage and current can be represented to a close approximation by harmonic functions whose wavelengths are integral multiples of $\pi$, the hyperbolic terms being negligible. In case (*b*) the spatial fundamental frequency varies within wide limits. For values of $\lambda > 1$, the fundamental frequency is to a first approximation equal to $\pi/2$, but this does not hold if $\lambda < 1$, when the hyperbolic terms must also be taken into account.

621.319.45.025 1100

**Study of Alternating-Current Electrolytic Capacitors**—W. C. van Geel and A. J. Dekker. (*Philips Res. Rep.*, vol. 5, no. 4, pp. 250–261; August, 1950. In French.) Capacitors consisting of two Al plates coated with $Al_2O_3$ and immersed in an electrolyte are investigated by analogy with a system comprising two ideal capacitors in series, independently shunted by opposed rectifiers, also in series. The potential differences between the various parts of the actual system are then considered, taking into account the effect of the leakage current. Measurements confirming the theory are reported and oscillograms of voltages and currents are shown.

621.385.3:621.315.592:518.4:621.396.615
+621.396.645 1101

**Graphical Analysis of Transistor Characteristics**—L. P. Hunter. (Proc. I.R.E., vol. 38, pp. 1387–1391; December, 1950.) Graphical methods are described for determining the required values of circuit components of voltage amplifiers, current amplifiers, and oscillators, from the dc characteristics of the transistors used.

621.392 1102

**A Note on the Synthesis of Resistor-Capacitor Networks**—C. Belove. (Proc. I.R.E., vol. 38, pp. 1453–1454; December, 1950.) Comment on 1605 of 1950 (Bower and Ordung).

621.392 1103

**Wide-Band Two-Phase Networks**—H. J. Orchard. (*Wireless Eng.* vol. 28, No. 328, p. 30; January, 1951. Comment on 56 of February (Johannesson). It is suggested that a still better way of treating problems of wide-band networks is to reverse the Landen transformation and work from a suitable hyperbolic function. The advantage of the method is that the larger the bandwidth ratio, the easier does the computation become, only one step being necessary for bandwidths exceeding one decade.

621.392 1104

**Wide-Band Two-Phase Networks**—W. Saraga. (*Wireless Eng.* vol. 28, no. 328. pp. 30–31; January, 1951.) In recent papers by Darlington (1359 of 1950), Orchard (1356 of 1950), and Saraga (2,736 of 1950) on wide-band networks, elliptic functions are used for obtaining a Tchebycheff approximation to the ideal performance. The expressions given for this approximation differ very much in form, although defining the same mathematical relation. Their equivalence is here demonstrated by reference to certain standard formulas in the theory of elliptic functions and their transformation.

621.392:621.387.424 1105

**Auxiliary Electronic Circuits for Geiger Counters**—L. Fontes and R. Moret. (*Radio franç.*, pp. 5–15 and 18–24; November and December, 1950.) Description of quench circuits, amplifiers, stabilized hv supply units, scaling-down circuits, including decade systems, and integrating circuits.

621.392.4/.5 1106

**Selective Generation, Amplification or Rejection of Alternating Voltages without the aid of Oscillatory Circuits**—H. Pieplow. (*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 6, pp. 1031–1039; 24th December, 1949. In German.) Starting from the conditions which must be satisfied by feedback networks, rules are established for the construction of oscillator, amplifier, or rejector networks not including *LC* circuits. Examination of the way in which different couplings can be used shows that the sharpness of resonance or rejection of such networks tends toward a lower limit which is appreciably higher than in the case of oscillatory circuits and can only be lowered by special means not dependent on the circuit arrangement. It follows that complex circuits are especially not worthwhile when the frequency must be controllable.

621.392.4 1107

**A Comparison of the Bandwidths of Resonant Transmission Lines and Lumped *LC* Circuits**—A. van Weel. (*Philips Res. Rep.*, vol. 5, pp. 241–249; August, 1950.) The ratio of the bandwidths obtainable in the two cases was calculated and found to depend on the product $\omega Z_0 C$ ($\omega_0$=resonance frequency; $Z_0$=characteristic impedance of line; C=terminal capacitance.) The bandwidth of Lecher lines is always less than that of an *LC* circuit; the longer the line, the less the bandwidth. For a short-circuited $3\lambda/4$ line the bandwidth is at least four times less than for an *LC* circuit. The bandwidths of $\lambda/2$ and $3\lambda/4$ lines can be increased to that of a $\lambda/4$ line by introducing discontinuities in the characteristic impedance at certain points. A qualitative explanation of these effects is given which applies also to cavity resonators.

621.392.41 1108

**Impedance Synthesis without Use of Transformers**—R. Bott and R. J. Duffin. (*Jour. Appl. Phys.*, vol. 20, p. 816; August, 1949.) A method is outlined for the synthesis of an arbitrary impedance by series-parallel combinations of inductors, resistors, and capacitors.

621.392.41 1109

**Impedance Synthesis distributing Available Loss in the Reactance Elements**—W. Nijenhuis. (*Philips Res. Rep.*, vol. 5, pp. 288–302, August, 1950.) A study of the synthesis of 2-pole networks whose elements possess maximal losses. Difficulties arising from mutual inductances are overcome in the way suggested by Bott and Duffin (1108 above). It is shown how losses can be introduced into coils to simulate skin effects or eddy currents. A numerical example is worked out.

621.392.5 1110

**Influence of Losses on Attenuation of Four-Terminal Impedance Networks**—S. Brimberg. (*Kungl. tekn. Högsk. Handl., Stockholm*, no. 43, 30 pp; 1950. In English.) The influence of losses on complex image attenuation in the case of infinite idealized image attenuation and at band edges is calculated. The effect of losses on complex effective attenuation at band edges in symmetrically designed filters with symmetrical loads is also investigated and theoretical and experimental results are compared for a low-pass filter.

621.392.52 1111

**Attenuation Equalization within the Pass Band of Ladder Filters**—W. Wolman. (*Arch. Elektrotech.*, vol. 40. no. 1, pp. 30–36; 1950.) Analysis shows that the frequency-response curve of each section of a filter can be flattened by introducing small resistances in shunt or in series, effectively forming auxiliary meshes with complementary response curves. Measurements are reported in confirmation of the theory, and the effect of the compensation is illustrated in response curves for a band-pass filter of 6 half sections and for a low-pass filter of 4 sections. Factors affecting the choice of location of the auxiliary resistors, a matter of great practical importance, are discussed.

621.392.52 1112

**Unusual Ladder Filter. Applications in Audio and Radio Circuits**—F. G. G. Davey. (*Wireless World*, vol. 57, pp. 31–34; January, 1951.) The properties are discussed of a class of filters in which the impedances looking forward and backward from the central element bear a constant ratio to each other with change of frequency and maintain equal phase angles. The application is described of adjustable low-pass filters of this type between the speech coil of a loudspeaker and the corresponding transformer winding, and of band-pass versions for use as variable coupling elements at intermediate frequency.

621.392.5:621.395.44 1113

**Loss-Compensated Filters in Carrier-Frequency Systems**—H. Lehmann. (*Fernmeldetech. Z.*, vol. 3, pp. 415–417; November, 1950.) Discussion of the advantages of using asymmetric filters in ssb carrier systems, particularly in eliminating the need for lf equalizers, in affording a choice of design, and in improving stability. The discussion is based on experimental results for channel filters covering the range 12 to 24 kc.

621.396.611.3 1114

**Rolling-Ball Analogue of Coupled Circuits**—P. C. Magnusson. (*Elec. Eng.*, vol. 69, p. 1079; December, 1950.) Summary of American Institute of Electrical Engineers 1950 Fall General Meeting paper. The system consisting of two inductively coupled *LC* loops with negligible resistance has as dynamic analogue a ball rolling freely on a specially curved surface. The time integrals of the loop currents are used as displacements along horizontal co-ordinates $x_1$, $x_2$, with the loop currents themselves analogous to the velocity of the ball along those co-ordinates. The magnetic-field energy is analogous to the kinetic energy of the ball if the scales of $x_1$ and $x_2$ are proportional to the square roots of the corresponding self-inductances and $x_1$ is inclined to $x_2$ at an angle $\beta$, the cosine of which is equal to the coefficient of coupling. The height of any point of the surface, and hence the potential energy of the ball, is proportional to the dielectric field energy, contours of which are concentric ellipses. The ball path is shown which corresponds to initial circuit conditions where both capacitors are charged and circuit currents are zero. The method of deriving an analogue should be applicable to any dynamically conservative electrical or mechanical system with two degrees of freedom, and should be useful in the interpretation of transients in such systems.

621.396.611.31.001.11 1115

**Theory of Oscillations in Coupled Electromagnetic Cavity Resonators**—E. Ledinegg and P. Urban. (*Acta Phys. austriaca*, vol. 2, no. 2, pp. 198–213; September, 1948. In German.) General formulas are derived by a perturbation method for the frequencies of a system of cavity resonators coupled by similar resonant members. Except for the assumption of weak coupling between the resonators, which in the undisturbed state must give rise to at least one common frequency, no essential limitations are introduced in the calculations. The theory shows that the frequencies of such coupled systems satisfy a secular determinant, whose members represent energy terms determined by the individual elements of the cavity-resonator system, and which can accordingly be calculated. The theory is applied to the practical cases of (*a*) two concentric transmission-line sections with a coaxial coupling unit, (*b*) two inductively coupled quasistationary circuits.

621.396.615 1116

**The Self-Blocking Oscillator and its Applications**—J. Moline. (*Radio franç.*, pp. 11–15 and 12–16; December, 1950 and January, 1951.) The development of the blocking oscillator from one having a sinusoidal wave form, by increasing the coupling between the anode and grid circuits, is described and the blocking action is explained; the effects of the different circuit elements being considered separately. Various typical practical circuits are shown and their particular advantages are mentioned. Applications described include pulse and sawtooth generators, use in counter circuits, and so forth.

621.396.615:621.316.726:621.396.4 1117

**Stabilized Master Oscillator for Multichannel Communication**—E. W. Pappenfus. (*Electronics*, vol. 23, pp. 108–113; December, 1950.) Particular attention is paid to systems providing a wide range of operating frequencies but utilizing only a single reference crystal. A commercially available oscillator covering the band 2 to 4.5 mc is described. Using a 100 kc reference crystal, the oscillator provides an output of any desired frequency in the above range stabilized to within $\pm 5 \times 10^{-6}$ for wide temperature and humidity variations. Details of the servo-motor afc circuit are given and discussed. Any one of ten preset output frequencies can be selected by operation of a remote switch.

621.396.615:621.316.726.078.3 1118

**Frequency Stabilization of Oscillators**—J. Zakheim. (*Radio franç.*, pp. 1–4; November, 1950.) If the effective output impedance $Z$ of the amplifier section is given by $Z=R+jX$ and the transfer coefficient $k$ of the limiter by $k=\alpha+j\beta$, the phase angle is theoretically constant and independent of the over-all slope of the maintaining amplifier when $\beta\alpha=-XR$. Practical application of the formulas derived is illustrated by numerical calculation for an oscillator of the Wien-bridge type operating at 3 kc.

621.396.615.14.029.63 1119

**Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits**—Ratheiser. (*See* 1280.)

621.396.615.141.2 1120

**Some Aspects of Split-Anode Magnetron Operation**—H. J. Reich, J. C. May, J. G. Skalnik, and R. L. Ungvary. (Proc. I.R.E., vol. 38, pp. 1428–1433; December, 1950.) Resonators and tubes at present available are unsuitable for use in wide-band magnetron oscillators using butterfly-type resonators, but they could be more successfully used if specially modified for the purpose. The negative slope of part of the anode-current voltage characteristic, explainable by back heating, is absent when the current is varied rapidly. The best method of anode-current stabilization is that in which the magnetic field is produced by an electromagnet excited by the direct anode current.

621.396.615.17:621.385.832 1121

**A Simple Slow-Running Timebase**—C. J. Dickinson. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 505–506; December, 1950.) A detailed description of a circuit designed to overcome inherent disadvantages of earlier timebases. The circuit requires less critical adjustment and avoids the need for a large negative supply voltage.

621.396.645:621.3.018.78† 1122

**Universal Design Curves for Intermediate-Frequency Amplifiers with Particular Reference to Phase and Amplitude Distortion and their Compensation**—W. Hackett. (Proc. I.R.E., vol. 38, pp. 1408–1417; December, 1950.) Graphs and abacs are provided which enable the phase and amplitude distortion to be determined for various types of coupling network, including compensation combinations. Graphical methods are indicated for assessment of performance in terms of the product of gain and bandwidth, and for the interpretation of equation parameters in terms of circuit parameters.

621.396.645.029.3:621.385.3/.4 1123

**A Comparison of Triodes and Beam-Tetrodes as Power Output Valves in Audio Amplifiers**—D. Bruckmann, W. S. Carey, and D. J. Fuller. (*Trans. S. Afr. Inst. Elect. Eng.*, vol. 41, pp. 258–266. Discussion, pp. 266–267; September, 1950.) Tests made on balanced push-pull amplifiers show that the curve representing the change of harmonic distortion with increase in load resistance is parabolic for beam tetrodes, while for triodes it resembles a rectangular hyperbola. With reactive loading the increase in distortion is greater for the tetrode. As the load resistance rises, the intermodulation distortion for triodes drops to a negligible value, while for tetrodes it increases rapidly. The effects with reactive loading are similar to those observed in the case of harmonic distortion. The relative merits of triodes and tetrodes for use in af amplifiers are summarized.

621.396.645.371 1124

**Amplifiers with Selective Negative Feedback**—H. Boucke and O. Schmidt. (*Fernmeldetech. Z.*, vol. 3, pp. 417–420; November, 1950.) Basic circuits are described for suppressing either a harmonic or the fundamental in the output of an amplifier by incorporating a resonant circuit in the feedback line. Turning this circuit to the first harmonic reduces the distortion appreciably. A typical attenuation curve for a frequency-multiplier circuit shows a drop of 28 db in output level for the 800 cps fundamental below that obtained with a non selective cathode circuit.

621.396.662.085.4 1125

**Bandspreading and Scale Equalization for RC Tuning Networks**—C. F. Van L. Weiland. (*TV Eng.*, vol. 1, pp. 12–15, 28–30 and 21, 28; July, August, and October, 1950.) A continuous-selector type of circuit is described. By use of suitable tuning elements in series and parallel, a threefold scale elongation is achieved, with increased spread at the hf end of the scale. Circuits providing partially equalized and completely equalized bandspreading are shown and design formulas given.

621.396.69+621.38]:061.5(058.7) 1126

**Alphabetical Listings of All Components, Complete Units, Allied Products, used in Electronic Equipment for All Purposes**—(*Electronics, Annual Buyers' Guide Issue*, vol. 23, pp. D1-D164; Mid-month June, 1950.) A list giving the names of more than 2,500 manufacturers of electronic components and equipment. Products are arranged alphabetically under generic names, with subdivisions. A trade-name index and an index of manufacturers are also included.

## GENERAL PHYSICS

519.2:621.3.015.2 1127

**On a First-Passage-Time Problem**—F. L. H. M. Stumpers. (*Philips Res. Rep.*, vol. 5, pp. 270–281; August, 1950.) An investigation of the probability that a function starting at a time $t=0$ with a value $E_0$ will not exceed a value $E_1$ in the time interval 0-$t_1$. This probability is calculated (*a*) from the Fokker-Planck equation with suitable boundary conditions. (*b*) from an integral equation derived by Schrödinger. The theory has application to the pulse charging of a capacitor and also to fluctuation problems.

534.1+538.56]:517.93 1128

**On the Forced Vibrations of Quasi-Linear Systems**—Friedlander. (*See* 1177.)

534.26+[535.42:538.56 1129

**Theory of Diffraction at a Screen**—W. Franz. (*Z. Phys.*, vol. 128, pp. 432–441; 3rd October, 1950.) The author's successive-approximation method for diffraction calculations (3118 of 1949 and 323 of March) yields, for sharp-edged screens, functions which are singular at the edge and in the higher approximations give rise to divergent integrals. The singularities can be eliminated and the method made useful for higher approximations by assuming the integrands of the Kirchhoff integrals to fall to zero gradually, instead of discontinuously, beyond the edge. *See* also 2183 of 1950 (Braunbek).

535.42 1130

**A Note on the Diffraction of a Cylindrical Wave by a Perfectly Conducting Half-Plane**—P. C. Clemmow. (*Quart. Jour. Mech. Appl. Math.*, vol. 3, pp. 377–384; September, 1950.) The most easily derived solutions in diffraction theory are for the case of incident plane waves. The solution for a line source can then be obtained by a further integration. This is illustrated by the example of a line source in the presence of an infinitely thin, perfectly conducting half plane. The introduction of Hankel functions is avoided, and the solution appears in a useful form, equivalent to that given by Macdonald and analogous to Sommerfeld's solution for an incident plane wave.

537.52 1131

**Retroaction and Similitude Considerations in Relation to the Starting of the Electrical Gas Discharge**—W. Fucks. (*Arch. Elektrotech.*, vol. 40, no. 1, pp. 16–30; 1950.)

537.523.5:621.317.33.029.64 1132

**A Microwave Study of the High-Pressure Arc**—J. D. Cobine, E. P. Cleary, and W. C. Gray. (*Jour. Appl. Phys.*, vol. 21, pp. 1264–1267; December, 1950.) The Impedance of an arc at atmospheric pressure was measured at 1 kmc by means of a coaxial line and standing-wave detector. The arc reactance and resistance, essentially a lumped load at the end of the line, increase with arc length. The resistance decreases with increasing current and is approximately the same as the dc resistance. The reactance is capacitive and nearly independent of the current. Arcs in air, argon, and helium, with direct currents of 1 to 4 amps, were studied.

537.525:621.396.822 1133

**Noise Temperature of a D.C. Gas-Discharge Plasma**—S. Kojima and K. Takayama. (*Phys. Rev.*, vol. 80, no. 5, p. 907; December 1, 1950.) At 14 mc, the temperature of the plasma noise of discharge tubes containing argon at pressures of 1 mm and 3 mm Hg is found to be proportional to the electron temperature. The ratio of plasma-noise temperature to electron temperature is approximately 4.5:1 at 1 mm and 2:1 at 3 mm Hg. The noise temperature is nearly the same at 600 kc as at 14 mc, indicating that the noise observed is mainly thermal.

537.525.029.64:538.69 1134

**The Effect of Magnetic Field on the Breakdown of Gases at Microwave Frequencies**—B. Lax, W. P. Allis, and S. C. Brown. (*Jour. Appl. Phys.*, vol. 21, pp. 1297–1304; December, 1950.) Results are given of measurements of the electric field required to produce hf breakdown in helium at low pressure in the presence of a magnetic field of 0 to 3,000 gauss. Two relevant theories, the average electron theory and the Boltzmann theory, are presented, and the correspondence between them is discussed.

537.525.5 1135

**Techniques for Measuring the Dynamic Characteristics of a Low-Pressure Discharge**—B. T. Barnes and S. Eros. (*Jour. Appl. Phys.*, vol. 21, pp. 1275–1278; December, 1950.)

537.527 1136

**The Minimum Voltage and the Discharge Process on Flashover with Alternating Voltage in Humid Compressed Air**—H. Böcker. (*Arch. Elektrotech.*, vol. 40, no. 1, pp. 37–44; 1950.) Measurements were made of the Flashover voltage in air at pressures up to 24 atm with water-vapor content close to saturation, cylindrical glass insulators 0.6 to 3 cm long being inserted between the dished electrodes of the plane parallel gap. The lowest value of the reduced flashover voltage observed in the region of saturation is the "minimum voltage." Results are shown graphically and discussed.

538.3 1137

**Wheeler and Feynman's Theory of Electromagnetic Interaction**—O. Costa de Beauregard (*Rev. Sci.* (Paris) vol. 88, no. 3305, pp. 34–40; January to March, 1950.) Analysis in the light of other theories leads to criticism on the grounds that conditions for the irreversibility of the radiation process are contained in the postulates of the theory.

538.56:621.396.677 1138

**Reflection and Refraction of Microwaves at a Set of Parallel Metallic Plates**—F. Berz. (*Proc. IEE* (London), Part III, vol. 98, no. 51, pp. 47–55; January, 1951.) A mathematical analysis of the problem based on Maxwell's equations and on periodicity and continuity considerations. "Formulas are obtained giving the nature, direction, phase, and amplitude of the reflected and transmitted waves. The reflected waves follow the usual laws found in the grating theory. In the case when only one non-evanescent plane wave is reflected, the transmission and reflection power coefficients are those obtained at the junction of two semi-infinite transmission lines corresponding to the free space and the plate medium respectively. The analogy with transmission lines is more complex when two non-evanescent waves are reflected. For the usual angles of incidence and plate spacings, the transmission power coefficient is high (maximum 99.5 per cent), and the phase shift on transmission is small."

538.561 1139

**Radio-Frequency Emission due to the Gyromagnetic Effect in a Discharge**—M. Laffineur and C. Pecker. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231. no. 25, pp. 1446–1447; December 18, 1950.) The source of electrons used was a Penning vacuum gauge (1525 of 1937) with two cold cathodes, one on each side of a common ring-shaped anode. This was evacuated, placed in the field of an electromagnet producing fields up to 300 gauss, and a voltage of 2 kv was applied. The em waves emitted were detected by means of a receiver tuned to a wavelength of 55 cm and connected to a pickup loop at the side of the apparatus. The best results were obtained at a pressure of $10^4$ mm Hg. A curve shows the voltage at the receiver terminals, less the background noise (equivalent to about 50 μv at the input), for fields from 184.5 to 194.5 gauss. This curve has a very sharp peak at 193.5 gauss, the corresponding gyromagnetic frequency being precisely that for which the receiver was tuned. A much smaller peak at about 190 gauss is attributed to an image frequency of the receiving apparatus.

549.321.13:537.311.3:537.533.9 1140

**Equilibrium Currents induced in Zinc Blende by Electron Bombardment of Negative Electrode**—M. F. Distad. (*Phys. Rev.*, vol. 80, no. 5, pp. 879–886; December 1, 1950.)

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7:538.122 1141

**Visual Investigation of the General Magnetic Field of the Sun**—G. Thiessen. (*Z. Astrophys.*, vol. 26, no. 1, pp. 16–30; July 31, 1949.) The method of measurement and the results obtained during 1947 to 1948 with the 60-cm refractor of the Hamburg observatory are described. Measurements on various *Cr* and *Fe* spectral lines show that at the time of the observations the pole field strength in the direction given by Hale, could not have a value >5 gauss; the results rather indicate a very weak field in the reverse direction.

523.7+523.854]:621.396.822 1142

**Electromagnetic Waves radiated by Fast Protons in Strong Magnetic Fields, and Correlation between Cosmic Radiation and Radio Noise from the Sun and the Galaxy**—B. Kwal. *Compt. Rend. Acad. Sci.* (Paris), vol. 231, no. 20, pp. 1057–1059; November 13, 1950.) The classical theory of radiation from a charged particle in a magnetic field shows that, in fields of the order of $10^3$ to $10^4$ gauss, protons of energy $10^9$ to $10^{10}$ ev radiate meter and centimeter waves. Hence it seems possible to associate radio noise with the acceleration of cosmic rays in sunspots and in similar spots occurring in stars.

523.752:[551.510.535+537.591 1143

**The Chromospheric Flare of 19th November 1949 and its Geophysical Consequences**—R. Bureau and A. Dauvillier. (*Ann. Géophys.*, vol. 6, no. 2, pp. 77–103; April to June, 1950.) Close correlation was observed between ionospheric and cosmic-radiation effects associated with this flare. The rarity of such a correlation is emphasized. The two authors describe respectively their detailed investigations of the two effects. Observations from various parts of the world are reported and discussed, especially in relation to possible explanations of the delay of the corpuscular effects with respect to the electromagnetic effects. See also 1911 (Ellison and Conway), 2206 (Müller), and 3039 (Clay and Jongen) of 1950.

550.362:621.317 1144

**Theoretical Basis of a Method for Determining the Specific A. C. Resistivity of the Ground**—Weber. (*See* 1193.)

550.386/.387]:621.331:625.1 1145

**Disturbances of Terrestrial Magnetic Field and Currents by Electrified Railways**—G. Dupouy. (*Ann. Géophys.*, vol. 6, pp. 18–50; January to March, 1950.) Disturbances observed in the magnetograms recorded at Chambon la Forêt have been found to be caused by the Paris-Orléans line 28 km away. The effect is particularly noticeable when the continuity of the electric cables is broken, either accidentally or for purposes of maintenance. For horizontally stratified ground the disturbances can be calculated as a function of the leakage currents along the track. Agreement with measurements is satisfactory.

551.510.534/.535 1146

**Stratospheric-Ionospheric Relationships**—N. C. Gerson. (Proc. I.R.E., vol. 38, p. 1456; December, 1950.) Investigations reveal no correlation between either the temperature or pressure at the tropopause and the critical frequency of any of the ionospheric regions, and also no correlation between temperatures at a height of 13 km and critical frequencies. Some slight correlation was found in several instances between the pressure at a height of 13 km and the critical frequencies of the $E$ and $F_2$ regions, but no correlation with sporadic $E$ conditions.

551.510.535 1147

**Magneto-Ionic Triple Splitting of Ionospheric Waves**—O. E. H. Rydbeck. (*Jour. Appl. Phys.*, vol. 21, pp. 1205–1214; December, 1950.) A full theoretical discussion of the third magneto-ionic component, or *z* trace, which has been observed frequently at the new Kiruna observatory. Strong coupling often exists between the ordinary and *z* components at a critical level. The *z* wave becomes strongly excited when the collisonal frequency $\nu$ approaches the critical value $\nu_c$ of Appleton and Builder. When $\nu$ is larger than $\nu_c$, the ordinary echo rapidly disappears and only the *z* and extraordinary components remain. A useful expression for the *oz* wave transmission coefficient is derived. Transmission of appreciable strength is possible only at high latitudes, where $\nu_c$ is fairly small. Experimental results obtained at Kiruna (magnetic latitude 65°) are reported.

551.510.535 1148

**Preliminary Results of Ionospheric Observations made at Dakar**—F. Delobeau, E. Harnischmacher, and F. Oboril. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 17–20; January to March, 1950.) Discussion of records obtained during the period May to September, 1949. Monthly mean values of $f_0F_2$ conform to the general shape of curves plotted for comparable latitudes, but extreme values are more pronounced: the highest critical frequencies in May were estimated at about 15 mc; median midday values for June and July were 13.8–13.5 mc and the night value for June was 4.2 mc. This lends support for the revision of zone demarcation proposed by Oboril and Rawer (2794 of 1949). Strong diffusion effects observed cannot be correlated with magnetic disturbances. The presence of an intermediate layer between $F_1$ and $F_2$ is indicated; the true height of the $F_2$ layer deduced from virtual-height measurements may reach 500 km. Values of $k$ and $n$ in the formula $f_0E=k(\cos\chi)^n$ are tabulated. Records of $E_s$ layer characteristics are also shown and discussed.

551.510.535:523.745 1149

**Long-Term Relation between $F_2$-Layer Ionization and Solar Activity, over the Whole World**—R. Gallet and K. Rawer. (*Ann. Géophys.*, vol. 6, nos. 2 and 3, pp. 104–114 and 212–213; April to June and July to September, 1950.) Close correlation is found between the monthly mean value of the $F_2$ layer maximum electron density and the relative sunspot number, on comparing the running means over 13 months. The coefficients of the linear correlation equations for 12 stations in different parts of the world are plotted against, (*a*) geographic latitude, (*b*) geomagnetic latitude, and (*c*) magnetic inclination; the last method gives the most satisfactory picture of their systematic variation. Abnormal relations are observed at the magnetic equator; observations from stations in that region are reserved for separate study.

551.510.535:551.594.5 1150

**Ionospheric Disturbance caused by the Aurora Borealis of 20th–21st February 1950**—R. Eyfrig. (*Ann. Géophys.*, vol 6., pp. 70–73; January to March, 1950.) A direct observation of an aurora borealis, relatively rare in latitudes as low as 48°, was made at Freiburg, starting at 1845 universal time on February 20. Unusually high electron concentration in the $F_2$ layer preceded and accompanied the observation. Further abnormalities of the $E$ and $F$ layers are shown in a series of $h'f$ records for the period from 1830 to 2200. The disturbance is interpreted as a horizontally travelling dip in the ionosphere. $f_0F_2$ was low throughout the night and until the next afternoon. A marked ionospheric disturbance was also observed on the antarctic on board the *Commandant Charcot* on February 20.

551.510.535:551.594.51 1151

**New Results of Importance in the Study of Auroral Spectra and the Physics of the Ionosphere**—L. Vegard. (*Ann. Géophys.*, vol. 6, pp. 157–163; July to September, 1950.) Results more complete and more precise than ever before have been obtained at Oslo, using a new spectrograph for analysis of the aurora of the 23rd to the 24th of February, 1950. Fifty-four new bands or lines were observed. The presence of molecular oxygen was proved for the first time. Many lines are due to nitrogen and oxygen, neutral and ionized, as previously observed. The temperature of the luminous region was estimated to be −54° centigrade.

551.510.535:621.3.087.4 1152

**An Automatic Ionospheric Recorder for the Frequency Range 0.55 to 17 Mc/s**—Naismith and Bailey. (*See* 1191.)

551.594.11 (988) 1153

**Measurements of the Electric Field of the Atmosphere in Greenland between Sea Level and the Centre of the Ice-Cap**—P. Pluvinage and G. Taylor. (*Ann. Géophys.*, vol. 6, p. 69; January to March 1950.)

## LOCATION AND AIDS TO NAVIGATION

621.396.9:526.9 1154

**The Application of Radar to Geodetic Surveying**—J. Warner. (*Aust. Jour. Appl. Sci.*, vol. 1, pp. 133–146; June, 1950.) A modified shoran radar equipment is used for measuring distances between 160 and 310 miles to an accuracy within about 7 parts in $10^5$. The method of measurement, equipment, used reduction of observations, and errors involved are described. The greatest errors were instrumental; it is considered that modifications of the equipment would reduce them to about 2 parts in $10^5$.

621.396.9:551.578.1/.4 1155

**Radar Echoes from Meteorological Precipitation**—J. E. N. Hooper and A. A. Kippax. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 303–394; November, 1950.) Discussion of paper abstracted in 2222 of 1950.

621.396.9:621.396.11 1156

**Some Adverse Influences of Meteorological Factors on Marine Navigational Radar**—Saxton and Hopkins. (*See* 1228.)

621.396.932.1:621.396.677 1157

**Cheese Aerials for Marine Navigational Radar**—Kiely, Collins, and Evans. (*See* 1080.)

621.396.933.2 1158

**Rho-Theta System of Air Navigation**—P. C. Snadretto. (*Elec. Commun.* (London), vol. 27, pp. 268–276; December, 1950.) Description of equipment designed to fulfil I.C.A.O. recommendations for short-range navigational aids. Azimuth (theta) is obtained by reception of signals from a ground vhf transmitter radiating (*a*) a carrier with 30 cps tone modulation from an antenna having a circular horizontal radiation pattern and (*b*) an unmodulated carrier of the same frequency radiated from a dipole rotating at 1,800 rpm. The phase relations of the two carriers remain rigidly fixed. Azimuth is obtained from the measurement of the phase difference between the tone from the rotating system and that from the omnidirectional system. Distance (rho) is determined from the time taken by a pulse radiated from the aircraft at uhf to travel to a ground station and be returned, at a different uhf, to the aircraft.

## MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.12.088 1159

**Reading Errors with the McLeod Gauge**—P. Romann. (*Le Vide*, vol. 3, pp. 522–530; November, 1948.)

533.56 1160

**Study and Realization of a New Rotary Molecular Vacuum Pump**—H. Gondet. (*Le Vide*, vol. 3, pp. 513–521; November, 1948.) Description of a pump of the Holweck type in which a disk is used instead of the usual drum, so that an increased output is obtained. The optimum dimensions for the various channels are calculated and pumping rates are given for an experimental pump with a duralumin disk, of diameter 40 cm and thickness 1.2 cm, rotating between two duralumin castings with spiral channels of cross section rapidly decreasing toward the center. Clearance between the disk faces and the partitions separating the channels is of the order of 0.03 mm. The driving motor is fitted within the part of the machine connected to the backing pump.

537.311.33:546.289 1161

**Pressure Dependence of Resistance of Germanium**—J. H. Taylor. (*Phys. Rev.*, vol. 80, pp. 919–920; December, 1950.) The application of hydrostatic pressure increases the width of the forbidden band and hence increases the resistivity. Measurements, at pressures up to 4,500 pounds per inch$^2$, of the change of resistivity with pressure for several samples leads to the value $10.2\pm0.4\times10^5$ atm for the pressure coefficient of resistance at 300° K. The temperature rate of change of width of the forbidden band then follows as $0.87\times10^4$ ev° Centigrade.

537.311.33:546.289:539.164.9 1162

**Changes in Conductivity of Germanium Induced by Alpha-Particle Bombardment**—W. H. Brattain and G. L. Pearson. (*Phys. Rev.*, vol. 80, pp. 846–850; December 1, 1950.) The bombardment of $n$-type Ge by $\alpha$ particles first removes the conduction electrons at the rate of 78 per $\alpha$ particle. After these electrons are gone, conducting holes are introduced at the initial rate of 8.6 per $\alpha$ particle. There are 6.6 of these conducting holes that decay with a half life of 13 hours at room temperature and the remaining two holes are permanent at this temperature. Conductivity is changed only to the depth of penetration of the particles, namely $1.9\times10^3$ cm. The distribution of holes with depth is not uniform.

537.312.5:546.46.86 1163

**Photoconductivity in Magnesium Antimonide Layers**—T. S. Moss. (*Proc. Phys. Soc.*, (London), vol. 63, no. 372B, pp. 982–989; December 1, 1950.)(Layers of $Mg_3Sb_2$ evaporated in vacuo act as semiconductors and are photoconductive. Current to voltage and radiation intensity to voltage relations, response time, spectral dependence of photosensitivity, and the resistance to temperature relation were investigated. Threshold wavelengths of 1.5$\mu$, 2.6$\mu$ and 3.5$\mu$ were found, the first corresponding to intrinsic conductivity ($\epsilon=0.81$ev) and the latter two ($\epsilon=0.48$ev and 0.35ev) to impurity levels. Recent Russian work is in agreement with these results [*Zh. tekh. Fiz.*, 1949 vol. 18, pp. 1459–1477 (Boltaks and Zhuze)].

537.312.8:669.15 1164

**A Study of the Magneto-Resistance of Silicon-Iron**—R. Parker. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 996–1004; December 1, 1950.) Curves for $\Delta\rho,\rho$ against field strength were obtained for single crystals and polycrystalline speciemns, for both longitudinal and transverse magnetization. Strong fields give a linear decrease less than 0.02 times that for Ni and Co. Weak fields give unusual features. It is concluded that the small magneto resistance of polycrystalline silicon iron is due to the small contribution of individual crystallites and not to an averaging-out process.

537.312.8.029.63:546.87 1165

**Magnetoresistance of Bismuth at 3,000 Mc/s**—C. W. Heaps. (*Phys. Rev.*, vol. 80, no. 5, pp. 892–893; December 1, 1950.) Using a resonant re-entrant cavity machined from cast Bi, fed through a slotted coaxial waveguide, swr measurements indicated that at 3 kmc the magnetoresistance of Bi is not more than half the dc value. This is possibly due to the skin depth being of the same order as the mean free path of electrons in the metal.

538.221 1166

**Oxide Ferromagnetic Materials**—E. Flegler. (*Arch. Elektrotech.*, vol. 40, pp. 4–16; 1950.) An account is given of the experimental investigation of various $\nu$ iron oxides and ferrites subjected to alternating fields. The variations with frequency of the conductivity, permittivity, permeability, and loss factors are shown and conclusions are drawn regarding the causes of these variations. The different uses of the term "complex permeability" are briefly discussed.

538.221 1167

**High-Frequency Permeability of Ferromagnetic Materials**—F. V. Webster and K. S. Driver. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, p. 1040; December 1, 1950.) Preliminary results obtained with more accurate apparatus than that previously used [913 of April (Millership and Webster)] are reported. The general conclusions remain unaffected.

538.221:538.214 1168

**Investigations on the Reversible Susceptibility of Ferromagnetics**—R. S. Tebble and W. D. Corner. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1005–1016; December 1, 1950.) The reversible susceptibility $\kappa_r$ of long wire specimens was measured by applying small alternating fields and using a mutual-inducto-meter bridge. Results are discussed. The minimum contribution of reversible processes to the total change in magnetization varies from 10 per cent to 20 per cent for iron and nickel specimens. Gans' statement, that $\kappa_r$ is a unique function of intensity of magnetization and independent of magnetic history has definite limitations.

549.514.51:621.396.611.21 1169

**Quartz Crystals free from Harmonics**—J. J. Vormer. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 339–346; November, 1950.) Overtones of Y′ oscillations in X-cut crystals can be suppressed by shaping the width of the electrodes to a sine curve; this has the effect that the areas subjected to opposed fields are exactly equal for any harmonic. Experiments made to test the theory indicated that the intensity of all harmonics could certainly be reduced to less than 1 per cent of the fundamental. Alternatively, a field distributed sinusoidally in the Y direction may be used with ordinary electrodes to achieve the same effect.

621.314.634:537.311.33 1170

**Forming Processes in Thallium-Containing Selenium Rectifiers**—A. Hoffmann. (*Z. Phys.*) vol. 128, no. 3. pp. 414–431; October 3, 1950. A new method of measuring the capacitance of a dry rectifier, due to Lehovec, makes it possible to estimate the concentration of impurity centers at different production stages. For a certain Tl content of the counterelectrode there is (*a*) a reduction of concentration of impurity centers in the boundary zone, (*b*) an increase of path resistance, and (*c*) an increase of blocking power; (*a*) and (*b*) are quantitatively related. The influence of the Tl is markedly dependent on whether the counterelectrode is absent or present during the final heat treatment at 210° centigrade. The increase of blocking power cannot be explained simply by the reduction of the boundary field strength, but must be mainly due to other causes.

**621.318.2** **1171**
**Calculation and Construction of Permanent-Magnet Systems**—E. Steinort. (*Radio Tech.* (Vienna), vol. 26, pp. 545–548; November, 1950.) The design affording maximum flux in the air gap with minimum use of material is discussed. Formulas are given for optimum length and cross section of the magnet, the correction factor for length and the leakage flux as a function of air-gap inductance are shown graphically for ticonal and alnico materials. When the leakage characteristic of the material is not known, optimum length and cross section may be determined by an experimental method using an electromagnet.

**621.791.3** **1172**
**Solders and Soldering: Some Recent Advances**—H. C. Watkins. (*Metallurgia*, (Manchester) vol. 42, no. 254, pp. 372–376; December, 1950.) Survey of materials and techniques, with special reference to wiping solders, the soldering of aluminum, special solders for automatic processing, and flux-cored solders. The physical effects of surface roughening and of absorption of atoms of the liquid metal are briefly considered.

**666.1.037.5** **1173**
**Glass-to-Metal Seals**—J. A. Pask. (*Prod. Eng.*, vol. 21, pp. 129–134; January, 1950.) A discussion of the properties of various metals and special alloys used for sealing to glass. Only direct seals, with no intermediate material other than the metal oxide, are considered. Thermal expansion curves are given and the stresses arising from imperfect matching between metal and glass are analyzed. Other important factors are the electrical and thermal conductivities of the metal, and the thickness of the oxide film allowed to form on its surface.

**621.317.2:537.72** **1174**
**Techniques Générales du Laboratoire de Physique: Vol. II [Book Review]**—J. Surugue (Ed.). Publishers: Centre National de la Recherche Scientifique, Paris, and H. K. Lewis and Co., London, 336 pp., paper cover 42s., cloth 44s. (*Jour. Sci. Instr.*, vol. 27, p. 341; December, 1950.) Nine sections, contributed by different authors, deal respectively with: measurement of infrared radiation; technique of thin quartz fibers and Wollaston wires; casting thin cellulose films; electrometers; electrometer tubes and circuits; ionization counters; investigation of fields by means of electrolyte tanks; "micro-forge"; automatic regulation of temperature.

## MATHEMATICS

**512.831:518.6:681.177** **1175**
**Calculation of the Eigenvalues of Matrices by means of Punched-Card Machines**—P. Henrici. (*Z. angew. Math. Phys.*, vol. 1, pp. 185–189; May 15, 1950.)

**517.9** **1176**
**Natural Eigenvalue Problems**—E. Stiefel and H. Ziegler. (*Z. angew. Math. Phys.*, Vol. 1, pp. 111–138; March 15, 1950. Eigenvalue problems, rising from questions of stability or in connection with oscillations may be approached (*a*) as problems of the calculus of variations, (*b*) by establishing their differential equations, (*c*) by means of their integral or integrodifferential equations. Up to the present, method (*b*) has been developed further than the others, but it has the disadvantage that it is not applicable to problems whose eigenvalue is explicitly involved in the boundary conditions. To overcome this difficulty, it is advisable to use method (*a*) which, from a physical point of view, has the advantage of facilitating the mathematical interpretation of mechanical restrictions. The authors intend to generalize the theory of eigenvalues and to establish relations between the three methods for the general case of so-called "natural" problems. In the present paper the relation between methods (*a*) and (*b*) is established for the case of one independent and one dependent variable.

**517.93:[534.1+538.56** **1177**
**On the Forced Vibrations of Quasi-Linear Systems**—F. G. Friedlander. (*Quart. Appl. Math.*, vol. 3, part 3, pp. 364–376; September, 1950.) Discussion of the differential equation: $\ddot{x}+\omega^2x=kf(x, \dot{x}, t)$ where $\omega$, $k$ are parameters, $k$ being small, and $f(x, \dot{x}, t)$ is periodic with period $2\pi$ in $t$. Using an approximate solution due to Kryloff and Bogoliuboff, it is shown that, if $f(x, \dot{x}, t)$ is a polynomial in $x$, $\dot{x}$, $\cos t$, and $\sin t$, the amplitude $(x^2+\dot{x}^2/\omega^2)1$ settles down to an asymptotic range of variation of order $k$ as $t\rightarrow\infty$ which depends on the initial conditions, unless $\omega$ differs only by order $k$ from any one of a set of critical rational values. For these values, subharmonic resonance occurs. In this case the upper limit for the asymptotic range of variation of the amplitude is of order $k/|\omega\text{-}\omega_0|$, where $\omega_0$ is the critical frequency nearest to $\omega$. The extension of the argument to an equation in which $f(x, \dot{x}, t)$ is not a polynomial in $x$, $\dot{x}$, $\cos t$, and $\sin t$, but still periodic in $t$, is indicated. The asymptotic behavior of the phase $\tan^1 (x\omega/\dot{x})$ is also considered briefly. The reduction of the case of subharmonic resonance to an auxiliary differential equationof the first order is outlined, if either $\omega_0=m/n$ where $m$ and $n$ are large relatively prime integers, or $k/|\omega\text{-}\omega_0|$ is small, the results obtained in this way are in agreement with those obtained when subharmonic resonance does not occur.

**517.942.82:517.942.6/.9** **1178**
**Determination of Eigenvalues using a Generalized Laplace Transform**—M. D. Friedman. (*Jour. Appl. Phys.*, vol. 21, pp. 1333–1337; December, 1950.) The classical eigenvalue problems of mathematical physics are solved by means of a Laplace transform extended, not over the interval (o ∞) but over the interval of interest for the differential equation. The method is applied to the Hermite, Laguerre, and Bessel equations and to the equation for the hypergeometric polynomials which include the Legendre, Tchebycheff, and Jacobi polynomials as special cases.

**519.272.119:534.78** **1179**
**Correlation Function Analysis**—L. G. Kraft —(*Jour. Acous. Soc. Amer.* vol. 22, pp. 762–764; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. Discussion of theory on which the MIT digital electronic correlator is based, and presentation of speech correlation curves obtained with it.

**519.272.119:621.39.001.11** **1180**
**The Auto-Correlation Function**—D. A. Bell. (*Wireless Eng.*, vol. 28, no. 328, pp. 31–32; January, 1951.) An explanation is given of what the auto-correlation function is and what it can be used for, and a list of papers and books bearing directly on the subject is included.

**681.142** **1181**
**Electronic Computers: General Design and Construction**—L. Kosten. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 313–338; November, 1950.) An outline is given of the main parts of digital computers, that is, the input and output units, the memory, the arithmetic unit, and the external control. Magnetic, supersonic, and electrostatic memory systems are discussed Two adding units are described, working respectively on a parallel and on a series basis. A simplified block diagram of the external control of the EDSAC is shown, together with a schedule of orders.

**681.142** **1182**
**Electronic Computers: Mathematical Bases**—I. Boxma. (*Tijdschr. ned. Radiogenoot.*, vol. 15, pp. 299–310; Discussion, p. 311. November, 1950.) An examination is made of the conflicting considerations imposed in the choice of a calculating system by considerations of accuracy, speed and scope on the one hand and cost on the other. The advantages and disadvantages of the binary digital system are discussed, and the elementary mathematical operations required in that system are described.

**681.142** **1183**
**Special Devices for Differential Analyzers**—A. C. Cook, L. K. Kirchmayer, and C. N. Weygandt. (*Elec. Eng.*, vol. 69, p. 1080; December, 1950.) Summary of American Institute of Electrical Engineers 1950 Fall General Meeting paper. The range of the General Electric Company's differential analyser is being extended by the development of special devices for the generation of auxiliary functions, thus effectively increasing the number of integrators available for the solution of a given problem. Recent additions include, (*a*) a gear-ratio selector, (*b*) a remote indicator, (*c*) a transportation-lag device, (*d*) a recording counter, (*e*) dead-band units, (*f*) disconnect units; their functions are described.

**681.142** **1184**
**The Physical Realization of an Electronic Digital Computor**—A. D. Booth. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 492–498; December, 1950.) Details are given of circuits designed to convert an existing relay computer with magnetic storage (3179 of 1949) to fully electronic operation. The units discussed are the memory control, the digit amplifier, the shifting register, the magnetic gate circuits, and the adding equipment.

**681.142** **1185**
**Marginal Checking as an Aid to Computer Reliability**—N. H. Taylor. (PROC. I.R.E., vol. 38, pp. 1418–1421; December, 1950.) The presence of deteriorating components, which are liable to fail and cause errors in a digital computer, can be detected by a variation of circuit voltages, thus inducing actual failure, while a test program is performed. The failing components can then be systematically isolated. Trials have shown the value of this "preventive maintenance" in improving reliability.

**681.142:519.272.119:621.39.001.11** **1186**
**A Digital Electronic Correlator**—Singleton. (*See* 1237).

**681.142:621.383** **1187**
**Photoelectric Analog Computer**—E. C. Koenig. (*Electronics*, vol. 23, pp. 124, 126; December, 1950.) Description of equipment, making use of units comprising a single lamp for excitation of ten photocells arranged radially and having separate apertures and outputs. The photocells used are of the vacuum type with $AgC_{s_2}OC_s$ cathode. Advantages of the computer are enumerated.

## MEASUREMENTS AND TEST GEAR

**621.3.018.41(083.7):621.396.615.17/.18+621.317.761** **1188**
**Frequency Generation and Measurements**—W. S. Mortley: H. J. Finden. (*Electronic Eng.* (London), vol. 22, No. 274, p. 531

December, 1950. Comment on 1961 of 1950 and author's reply.

621.3.018.41(083.74)+621.316.726.078.3 1189
**The Evolution of Frequency Control**—Booth. (*See* 1271).

621.3.018.41 (083.74):621.396.81 1190
**Coverage of Standard-Frequency Station WWVH**—E. L. Hall. (*TV Eng.*, vol. 1, pp. 16–18 and 20–27; August and October, 1950.) A report of reception tests at several dozen stations in The United States, Alaska, Australia, China, Japan, and islands in the Pacific Ocean. The results are plotted and the resulting curves show reception efficiency as dependent on signal frequency, time of day, season of year, and sunspot conditions. The times of sunrise and sunset at the receiving station and at the transmitter have a definite effect on the percentage usability of the signals; the best reception is obtained for an all-darkness path. This is particularly the case for the 5-mc signals. The 10-mc signals are usable for more time than the 5-mc signals, while the 15-mc signals are usable still longer and may give uninterrupted service throughout the 24 hours at distances up to nearly 4,000 miles. The results in general indicate that for Alaska and the Pacific areas the transmissions from WWVH have resulted in an improvement of 87 per cent in the standard-frequency and time-signal service, though in The United States some confusion or interference with WWV transmissions, amounting to 2.5 per cent, has been caused.

621.3.087.4:551.510.535 1191
**An Automatic Ionospheric Recorder for the Frequency Range 0.55 to 17 Mc/s**—R. Naismith and R. Bailey. (*Proc. IEE* (London), Part III, vol. 98, pp. 11–18; January, 1951.) A full description of automatic equipment giving a photographic record of ionospheric characteristics in the form of a graph of height of reflection at vertical incidence against transmitted frequency. A pulsed self-oscillator coupled directly to the transmitting antenna is tuned from 0.55 to 17 mc in five bands. The receiver tuning capacitor is driven by a magslip motor whose speed is controlled by a discriminator circuit so as to keep the receiver in tune with the transmitter. The height and frequency calibrations and the date and time are recorded automatically on each record. The equipment has been in service for six years and has recorded 98 per cent of the hourly values of the critical frequency of the $F_2$ layer.

621.3.087.4:551.510.535 1192
**A Single-Band 0-20-Mc/s Ionosphere Recorder Embodying Some New Techniques**—T. L. Wadley. (*Proc. IEE* (London), Part III, vol. 98, pp. 45–46; January, 1951.) Discussion on 392 of 1950.

621.317:550.362 1193
**Theoretical Basis of a Method for Determining the Specific A.C. Resistivity of the Ground**—H. Weber. (*Tech. Mitt. schweiz. Telegr.-Teleph. Verw.*, vol. 28, pp. 257–260; July 1, 1950. In French and German.) Description of a method, similar to that of Collard (1136 of 1936), making use of a family of curves for the magnetic field produced by current in an overhead line with earth return.

621.317.3:621.392.5 1194
**Automatic Transmission Measuring Set**—J. M. Hudack. (*Bell Lab. Rec.*, vol. 28, pp. 538–541; December, 1950.) An outline description of the equipment and its principles of operation. It provides, in about 30 seconds, a record of the amplitude transmission characteristic of the apparatus under test as a function of frequency within the alternative bands 20 cps to 20 kc or 100 cps to 100 kc. Portions of these bands may be selected if required. Other uses for the equipment include the measurement of harmonic generation within the equipment under test and examination of its noise spectrum.

621.317.311+536.58:621.38+621.397.62 1195
**Electronics in Britain**—J. H. Jupe. (*N.Z. Elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The direct current or voltage to be measured is applied to a circuit including a short thin Pt wire whose resistance is varied periodically by application of hf current modulated at 50 cps. From the resulting fluctuations of the dc an alternating voltage is derived, amplified, and applied to an indicator calibrated in terms of direct current or voltage.

621.317.329 1196
**Rheographic Study of Laplace Fields with Helicoidal Structure**—H. Dormont. (*Philips Res. Rep.*, vol. 5, pp. 262–269; August, 1950. In French.) Two methods are described for determining the equipotentials of a spirally wound system of conductors from measurements in an electrolyte tank. *See* also 1206 and 1266.

621.317.334/.335.087.3:546.3 1197
**A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator)**—A. Hersping. (*Arch. Elektrotech.*, vol. 40, pp. 57–75; 1950.) The method is based on the phase variation within the pull-in range of a pair of coupled tube oscillators. A description is given of a metal locator embodying the principle. The theory of the pull-in action is given briefly; formulas derived are confirmed by experiments.

621.317.725.029.45/.5 1198
**Study of a H. F. Millivoltmeter**—(*Radio franç.*, pp. 1–6; December, 1950.) Description of a Philips instrument. *See* 2285 of 1950 (Lindenhovius and all).

621.317.755:621.3.011.6 1199
**Measurement of Time-Constants**—R. Aschen. (*Radio franç.*, pp. 16–19; November, 1950.) Description of the principle of operation of a cro circuit. Two EL41 tubes are connected in a multivibrator circuit which provides the horizontal sweep voltage and also controls the current through a third EL41, in the anode circuit of which is the component or circuit on test. When the tube is conducting, the scanning spot is held to the right of the screen; during flyback the tube anode current is cut off and the discharge characteristic of the component is displayed on the screen. Typical traces and complete circuit details are shown.

621.396.622.63.001.4 1200
**Mixer-Crystal Checker**—P. D. Strum. (*Electronics*, vol. 23, pp. 94–97; December, 1950.) A method for quick determination of conversion loss from dc measurements. The theory of the method is outlined and a practical instrument described; its accuracy is estimated by comparison with a series of direct measurements of conversion loss made at wavelengths of 3 and 10 cm. The basis of the method is that minimum conversion loss for a crystal can be predicted from its static voltage-current curve.

621.396.645.001.4 1201
**Test and Alignment Procedures for Video Amplifiers**—F. E. Cone and N. P. Kellaway. (*Broadcast News*, no. 61, pp. 28–33; September to October, 1950.) General procedures are outlined and specific alignment methods are described for each component of the RCA television-terminal equipment.

621.397.6.001.4 1202
**Signal Sources for Television Testing**—D. W. Thomasson. (*Jour. Brit. IRE*, vol. 10, pp. 369–390. Discussion, pp. 391–392; December, 1950.) A detailed discussion of the performance requirements of signal sources. The rf source must be within 0.1 per cent of the correct frequency. The video-frequency generator must give an output approximating to the actual television signal. Tolerances for the timing of the various parts of the synchronizing and picture signals are given. Suitable circuits are described for the complete generator.

621.317.3.029.6 1203
**Micro-Wave Measurements [Book Review]** H. M. Barlow and A. L. Cullen. Publishers: Constable and Co., London, 339 pp., 30s. (*Jour. Sci. Instr.*, vol. 27, p. 342; December, 1950.) A book of permanent value not only to communication engineers, for whom it is primarily intended, but also to research physicists. Clear explanations are provided of the principles involved in the measurements. The longest chapters are those dealing with standing-wave measurements and with matching and transmission systems.

621.396.6.001.4 1204
**Testing Radio Sets [Book Review]**—J. H. Reyner. Publishers: Chapman and Hall, London, 5th edn., 215 pp., 22s 6d. (*Jour. Sci. Instr.*, vol. 27, p. 341; December, 1950.) A practical book which can be recommended.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.316.79.076.7:621.369.3 1205
**Electronic Control of Home Heating**—J. M. Wilson. (*Electronics*, vol. 23, pp. 84–87; December, 1950.) Description, with complete circuit details, of an ac resistance bridge and associated amplifier controlling the heating system of a building, greater heat being switched on when the resistance variation of an outdoor element indicates a drop in temperature.

621.317.311+536.58:621.38+621.397.62 1206
**Electronics in Britain**—J. H. Jupe. (*N. Z. elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature control, and extension units for television.

The temperature-control unit operates by means of a Wheatstone-bridge circuit with a special diagonal branch including a resistor forming part of an amplifier circuit, a galvanometer, and two thermocouples connected in opposition whose action controls a servomechanism. *See* also 1195 and 1266.

621.317.334/.335.087.3:546.3 1207
**A Method for the Measurement of Small Variations of Inductance and Capacitance (Metal Locator)**—(*See* 1197).

621.365.44† 1208
**Load-Matching Dielectric Heaters**—R. H. Hagopian. (*Electronics*, vol. 23, pp. 98–101; December, 1950.) Increased efficiency is obtained by use of transmission lines for coupling rf generators to the loads, as matching problems are simplified.

621.38.001.8 1209
**Electronics Symposium and Exhibition**—(*Nature*, (London), vol. 166, no. 4234, pp. 1062–1063; December 23, 1950.) Summaries of the lectures at the opening ceremony of the conference arranged by the Scientific Instrument Manufacturers Association, London, September, 1950, and of the papers presented at the various sessions, with a note of the exhibits. For another account *see* 420 of March.

621.38.001.8:786.6 1210
**The Baldwin Electronic Organ**—A. Douglas. (*Electronic Eng.*, (London), vol. 22, no. 274, pp. 507–511; December, 1950.) An instrument in which the tones are generated entirely by tube oscillators. The basic waveform is a sawtooth, and square waves are generated by

the addition of two sawtooth waves. Instrument tones are simulated by the use of resonant filters. Details are given of the basic circuits and of various refinements designed to provide tone quality closely resembling that of a pipe organ.

621.383+621.385.38]:796.357 1211
**Thyratron-Controlled Photoelectric Umpire**—R. F. Shea. (*Electronics*, vol. 23, pp. 74–77; December, 1950.) Description of an automatic device that indicates passage of a baseball through the strike zone and also the speed of the ball as it crosses the plate. Three sets of photocells look at the sky through narrow slits, thus defining two vertical planes and an inclined plane intersecting the other two at knee and arm-pit height respectively. Passage of the ball through these planes in the correct sequence triggers an interconnected trio of thyratrons and operates an indicator.

621.384.611.2† 1212
**The Design of Electron Synchrotrons**—F. K. Goward, J. D. Lawson, J. J. Wilkins, and R. Carruthers. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 320–333. Discussion, pp. 334–339; November, 1950.)

621.384.611.2† 1213
**The Design and Operation of a 30-MeV Synchrotron**—D. W. Fry, J. Dain, H. H. H. Watson, and H. E. Payne. (*Proc. IEE* (London), Part I, vol. 97, no. 108, pp. 305–319. Discussion, pp. 334–339; November, 1950.)

621.384.612.1† 1214
**Beam Oscillations in an F. M. Cyclotron**—T. Teichmann. (*Jour. Appl. Phys.*, vol. 21, pp. 1251–1257; December, 1950.)

621.385.1.001.8:531.768.087 1215
**The Improved Ramberg Vacuum-Tube Accelerometer**—(*Tech. Bull. Nat. Bur. Stand.*, vol. 34, pp. 180–182; December, 1950.) Improved performance is obtained by, (*a*) an aging treatment resulting in more constant emission, (*b*) double gettering, (*c*) a 25-fold increase of sensitivity. See also 812 of 1949 and 2528 of 1947.

621.385.832:621.3.012:518.4 1216
**The Graph-Scope, an Electronic Graph Plotter and Graphical Computer**—A. L. Thomas, Jr. (*Elec. Eng.* vol. 69, pp. 1097–1100; December, 1950.) Description of an instrument for displaying on the screen of a cathode-ray tube a graph corresponding to any set of data, together with co-ordinate scales which can be either linear, logarithmic, or hyperbolic. A keyboard like that of a computer is used for setting the point-plotting circuits.

621.385.833 1217
**A New Ratiometer for the Electron Microscope**—R. Strauss. (*Arch. Elektrotech.*, vol. 40, pp. 49–56; 1950.) The object of this ratiometer is to make possible the full exploitation of the high resolving power of electron microscopes with magnetic lens systems, by accurately measuring and maintaining the ratio of the accelerating voltage to the square of the lens current, which ratio determines the image definition. The ratiometer comprises a long magnetic coil, within whose field the path of an electron beam is a helix; variations of the path are displayed on a fluorescent screen, and adjustment can be performed automatically or by hand. The accuracy attained for voltage variations is to within 2 parts in $10^4$.

621.385.833 1218
**Focusing in Electron Microscopy**—R. S. M. Revell and A. W. Agar. (*Jour. Sci. Instr.*, vol. 27, p. 337; December, 1950.) In high-resolution work, personal errors in judging optimum focus can be avoided by taking a through-focus series of micrographs.

621.385.833 1219
**The Correction of the Spherical Aberration of Electron Lenses Using a Correcting Foil Element**—U. F. Giancla. (*Proc. Phys. Soc.* (London), vol. 63, no. 472B, pp. 1037–1039; December 1, 1950.) A physical explanation of the action of foil elements, introduced into the active region of an electron lens, in correcting the positive aberration of the lens, is confirmed by rigorous analysis.

621.387.4† 1220
**An End-Window Alpha Scintillation Counter for Low Counting Rates**—C. W. Reed. (*Nucleonics*, vol. 7, pp. 56–62; December, 1950.)

621.387.424†:621.392 1221
**Auxiliary Electronic Circuits for Geiger Counters**—Fontes and Moret. (*See* 1105.)

778.37:537.523.4 1222
**A High-Intensity Short-Duration Spark Light Source**—J. A. Fitzpatrick, J. C. Hubbard, and W. J. Thaler. (*Jour. Appl. Phys.*, vol. 21, pp. 1269–1271; December, 1950.) Summary noted in 2604 of 1950 (Fitzpatrick and Thaler).

778.37:621.397.331.2 1223
**An Iconoscope Electro-Optical Shutter for High-Speed Photography**—H. A. Prime and R. C. Turnock. (*Proc. IEE* (London), Part II, vol. 97, No. 60, pp. 793–796; December, 1950.) The system described makes use of an iconoscope in which the photo-emission is controlled by the application of suitable voltages to the collector electrode. The charge image, formed during a selected interval, is stored on the mosaic and subsequently scanned. Examples of exposures of duration ranging down to 20 $\mu$s are shown. The mechanisms of spurious-image formation during the "shutter-closed" period are discussed. Performance is compared with that of the Kerr-cell type of shutter.

621.387† 1224
**Ionization Chambers and Counters [Book Review]**—D. H. Wilkinson. Publishers: Cambridge University Press, New York, N. Y., 1950, 255 pp., $4.50. (PROC. I.R.E., vol. 38, p. 1465; December, 1950.) The treatment is focused "upon the principles of instrumentation, particularly upon the Geiger counter and ion chamber, to the exclusion of any description of the crystal counter and some of the more recent nuclear detection devices. None the less, the book is very complete and well organized with respect to the more conventional counting devices."

## PROPAGATION OF WAVES

621.396.11+535.222 1225
**The Velocity of Propagation of Electromagnetic Waves derived from the Resonant Frequencies of a Cylindrical Cavity Resonator**—L. Essen. (*Proc. Roy. Soc. A*, vol. 204, no. 1077, pp. 260–277; December 7, 1950.) The length of the resonator was varied by mean of a piston whose change of position was measured by calibrated slip gauges. This change was measured for the $H_{01n}$ resonance modes in vacuo near 9 and 6 kmc up to $n=8$ and $n=2$, respectively. The frequency was measured by comparison with a quartz-crystal standard, using an interpolation oscillator. Measurements made at different frequencies allowed the diameter of the cavity to be determined without knowing $c$; the value obtained was greater by $2\times10^4$ cm than that determined by metrological methods; the difference is attributed to a film of AgS, which would also account for the 25 per cent difference between the measured and calculated values of $Q$.

From a survey of all possible sources of error the maximum estimated error is $\pm3$ km per second and the final value of $c=299792.5$ kms. *See* also 1751 of 1950, and 324 (Bergstrand) and 430 (Bol) of March.

621.396.11:551.510.535 1226
**The Path of a Ray in a Curved Ionosphere Layer**—K. Bibl. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 27–29; January to March, 1950.) Calculation of the effective ground-path length ($D_1$) for a reflected ray can be simplified by considering, not the angle of incidence on the lower surface of the layer, but the angle which the nonrefracted ray makes with the median plane of the layer, and by starting integration from this plane. The expression derived for $D_1$ indicates that the plane-layer formula is applicable in the case of a curved layer, the maximum error not exceeding 25 km. For nearly horizontal propagation the second-order (frequency correction) terms cannot be neglected.

621.396.11:621.396.812.4 1227
**Comparison of Tropospheric Reception at 44.1 Mcs with 92.1 Mc/s over the 167-Mile Path of Alpine, New Jersey, to Needham, Massachusetts**—G. W. Pickard and H. T. Stetson. (PROC. I.R.E., vol. 38, p. 1450; December, 1950.) Summary only of paper abstracted in 977 of May.

621.396.11:621.396.9 1228
**Some Adverse Influences of Meteorological Factors on Marine Navigational Radar**—J. A. Saxton and H. G. Hopkins. (*Proc. IEE* (London), Part III, vol. 98, pp. 26–36; January, 1951.) The effect of absorption and scattering of centimeter radio waves by atmospheric gases, uncondensed water vapor, and various forms of precipitation on the range obtained by marine radar equipment is discussed. For the wavelength of 3.2 cm now used for marine radar, the attenuation by atmospheric gases and uncondensed water vapour is hardly significant; absorption and scattering by rain are likely to cause most of the appreciable reductions in range of detection. It appears that, at a wavelength of 3.2 cm, for targets having echoing areas greater than 2,000 m$^2$ (that is ships of more than about 10,000 tons), attenuation in rainstorms will cause reduction in range, whereas for smaller targets masking by echoes from the rain will often be the more serious factor. Although very intense snowstorms can produce echoes sufficiently strong to be troublesome, the rate of precipitation required is such that its frequency of occurrence is unlikely to be great. In dense fogs reduction in detection range may be appreciable. Tables and graphs are given showing the effect on detection range of various types of precipitation in polar, temperate, and tropical regions for ships of 10,000 tons and of 1,000 tons, and for small boats or buoys.

621.396.11.029.62 1229
**The Propagation of Metre Waves**—F. C. Saic. (*Elektrotech. u. Maschinenb.*, vol. 67, pp. 325–332; November, 1950.) Critical discussion of the formulas of van der Pol and Bremmer and of Eckersley for propagation over land and over sea, both within and beyond the "visibility" range, taking account of transmitter height. Experimental results for wavelengths in the range 3 to 8 m are shown graphically.

621.396.81:621.3.018.41(083.74) 1230
**Coverage of Standard-Frequency Station WWVH**—Hall (*See* 1190.)

621.396.812 1231
**Radio Propagation between Noumea and Adélie Land**—M. Barre, K. Kawer, and E. Argence. (*Rev. Sci.* (Paris), vol. 88, no. 3305, pp. 21–26; January to March, 1950.) A series of graphs shows the relative strengths at various times of day of communication signals at different frequencies received at Noumea and in the steam ship *Commandant Charcot* about 5,500 km away in the antarctic. The optimum time for communication was about 1100 GMT (2100 local time), using frequencies in the range 8 to 17 mc. Conditions generally

remained good during the night but deteriorated abruptly at dawn. Factors used in predicting usable frequencies for the voyage are considered; due account was taken of auroral effects and of records formerly obtained at stations in the corresponding northern latitudes. The predictions show remarkable agreement with observations.

### RECEPTION

621.396.812 1232

**Diversity Effects in Spaced-Aerial Reception of Ionospheric Waves**—E. N. Bramley. (*Proc. IEE* (London), Part III, vol. 98, pp. 19–25; January, 1951.) A mathematical analysis is given of the phase and amplitude of the signals induced in two spaced antennas by a continuous angular spectrum of downcoming rays, all lying in the vertical plane through the two antennas. It is assumed that the amplitudes of the individual rays are constant and that their phases vary in a random manner. If the energy is concentrated in a small range of angles the extent of the angular distribution can be estimated from observations of the variation of either the amplitudes or the phase difference of the signals in the spaced antennas. A similar analysis is made for a specularly reflected component superimposed on an angular distribution of waves having random phase. If observations are made of both the amplitudes and the phase difference of the signals in the two antennas, the width of the angular distribution and the ratio of the amplitude of the specularly reflected component to that of the angular distribution (the signal noise ratio) may be found.

Some daytime measurements of first-order reflections from the ionosphere at nearly vertical incidence at frequencies between 4 and 7 mc are most satisfactorily explained in terms of a steady specularly reflected component, superimposed on a noise background. Values of signal noise ratio of about 2 or 3 have been obtained with a noise background having an angular spread of about 1 degree.

621.396.823 1233

**Car Ignition Radiation**—C. C. Eaglesfield. (*Wireless Eng.*, vol. 28, no. 328, pp. 17–22; January, 1951.) Theory previously given (3741 of 1946) for the radiation from car ignition systems is developed, assuming that the impulse voltage is applied to the sparking plug through a radiating inductor. Allowance is made for resonances in the ignition system and for the addition of suppression resistors. Satisfactory agreement is obtained with the experimental results of Pressey and Ashwell (1779 of 1949) for the latter case.

621.396.822 1234

**Threshold Signals [Book Review]**—J. L. Lawson and G. E. Uhlenbeck. Publishers: McGraw-Hill, New York, N. Y., 1950, 388 pp., 42s. 6d. (*Electronic Eng.*, vol. 22, No. 274, p. 532; December, 1950.) Volume 24 of the Radiation Laboratory Series. " . . . presents a connected account of the work at the Radiation Laboratory on the discrimination of signals in the presence of noise, with particular emphasis on pulsed signals, visually displayed in the presence of random noise. . . . The volume can be highly recommended both to the newcomer and to the specialist."

### STATIONS AND COMMUNICATION SYSTEMS

621.39:001.8 1235

**Storage Devices for Communications**—A. J. Lephakis. (*Electronics*, vol. 23, pp. 69–73; December, 1950.) The advantages to be gained by systematically recording messages prior to transmission over a communications link are considered. For such purposes storage systems are required in which the rate of extraction of data is not at all times equal to the input rate; possible systems are indicated. Other uses of storage devices considered include (*a*) radar display gear designed to respond only to moving targets and (*b*) equipment in which a desired relation between input and output signals (transfer characteristic) is synthesized.

621.39.001.11:519.272.119 1236

**The Auto-Correlation Function**—Bell. (*See* 1180.)

621.39.001.11:681.142:519.272.119 1237

**A Digital Electronic Correlator**—H. E. Singleton. (*Proc. IEE* (London), vol. 38, pp. 1422–1428; December, 1950.) 1950 IRE National Convention paper. "The relation between correlation functions and the general theory of communication is presented, and this relation leads to a technique for electronic computation of correlation functions and to the design of a machine for carrying out the computation. Because of the requirements of great accuracy and long storage, the machine makes use of binary digital techniques for storage, multiplication, and integration. Descriptions of the more unusual circuits in the machine are given, and circuit diagrams are included. A number of experimental results obtained by the machine are presented."

621.392.001.11 1238

**The Information-Theory Point of View in Speech Communication**—R. M. Fano. (*Jour. Acous. Soc. Amer.*, vol. 22, pp. 691–696; November, 1950.) A paper presented at the Speech Communication Conference at Massachusetts Institute of Technology, 1950. A nonmathematical discussion of the Wiener and Shannon theory of information, which is used to estimate the rate of transmission of information in speech communication.

621.395.44:621.392.52 1239

**Loss-Compensated Filters in Carrier-Frequency Systems**—Lehmann. (*See* 1113.)

621.396.4:621.396.615:621.316.726 1240

**Stabilized Master Oscillator for Multichannel Communication**—Pappenfus. (*See* 1117.)

621.396.41:621.396.619.13 1241

**The Application of Frequency Modulation to V.H.F. Multichannel Radiotelephony**—J. H. H. Merriman and R. W. White. (*Proc. IEE* (London), Part III, vol. 98, p. 56; January, 1951.) Discussion on 3513 of 1948.

621.396.619.13:621.3.018.42† 1242

**Bandwidth of a Sinusoidal Carrier Wave, Frequency-Modulated by a Rectangular Wave with Half-Sine-Wave Build-Up**—W. A. Cawthra and W. E. Thomson. (*Proc. IEE* (London), Part III, vol. 98, pp. 69–74; January, 1951.) "The problem of determining the frequency spectrum of a sinusoidal carrier, frequency modulated by a modified rectangular wave, is considered and a general solution obtained. The modification of the rectangular wave consists in replacing the vertical sides by a noninstantaneous transition in the form of a half-cycle sine wave. It is shown that for the determination of bandwidth it is reasonable to use, instead of the spectrum proper, a function which forms an envelope to it; this function is simpler and, in particular, it is independent of the on or off ratio of the rectangular wave. Curves are given for various values of the parameters involved."

621.396.65:[621.317.083.7+621.396.5 +621.398 1243

**Field Testing a Microwave Channel**—D. R. Pattison, M. E. Reagan, S. C. Leyland, and F. B. Gunter. (*Elec. Eng.*, vol. 69, pp. 1092–1097; December, 1950.) Essentially full text of AIEE 1950 Summer and Pacific General Meeting paper. A description is given of microwave channel and af multiplex terminal equipment installed by the Pennsylvania Electric Company for simultaneous relaying, telemetering, supervisory control, and voice communication. A reflector is used to provide a continuous microwave path between terminal stations where direct line-of-sight communication is not possible.

621.396.65:621.396.828 1244

**Reflected-Ray Suppression**—H. E. Bussey. (Proc. I.R.E., vol. 38, p. 1453; December, 1950.) Destructive interference between the direct and ground-reflected rays may impair line-of-sight microwave communication. The reflected signal may be removed by means of an opaque screen on the ground at the geometrical reflection point, which blocks out one half of the first Fresnel zone. Confirmatory tests for an 800 foot transmission path operating on a frequency of 4.5 kmc are described.

621.396.65.029.63 1245

**A Simplified Method of Planning Decimetre-Wave Line-of-Sight Links, taking into Account the Earth's Curvature and the Vision Ellipse**—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 3, pp. 408–414; November, 1950.) Description, with relevant curves and abacs, of a geometrical method based on constructing the ellipse representing the first Fresnel region. The method is applied to a 50-km link using a wavelength of 12.5 cm.

621.396.712 1246

**WBSM, WBSM-FM, New Bedford, Massachusetts**—O. F. A. Arnold. (*Broadcast News*, no. 61, pp. 9–15; September to October, 1950.) Illustrated description of studios, control room, and transmission system.

621.396.712 1247

**WKPT and WKPT-FM**—T. Phillips. (*Broadcast News*, no. 61, pp. 22–27; September to October, 1950.) Illustrated description of studios, control rooms, and transmission system at Kingsport, Tennessee.

621.396.712(94) 1248

**Engineering Aspects of the National Broadcasting Service**—R. J. Boyle. (*Telecommun. Jour. Aust.*, vol. 7, no. 1, pp. 16–30; June, 1948.) An outline of the establishment of the National Broadcasting Service in Australia, and discussion of the problems involved, and the methods used to solve them. The planning of the transmitter locations, site selection, and field-strength surveys with test transmitters, are first considered. Typical transmitter equipment is then described, with construction details of the top-loaded vertical radiators used at Cumnock, NSW, Dooen, Victoria, and Wagin, Western Australia. As standard equipment, all transmitting stations are provided with a beat-frequency oscillator, noise and distortion meter, and modulation monitor; all have stand-by power plant.

Practically all the program material broadcast from National stations originates in, or is routed through, the studios in the capital cities. A general account of the studio and program-switching arrangements is illustrated by reference to the Sydney studios. Relay systems for interstate and National broadcasting, and the arrangements at Liverpool, NSW, for the reception and rebroadcasting of overseas programs, are also described.

### SUBSIDIARY APPARATUS

621-526 1249

**Relay Servomechanisms: The Shunt-Motor Servo with Inertia Load**—T. A. Rogers and W. C. Hurty. (*Trans. Amer. Soc. Mech. Engr.*, vol. 72, pp. 1163–1172; November, 1950.) The theory is developed in terms of dimensionless motor parameters. Operating curves are drawn in the phase plane illustrating the effect of parameter changes on the stability of the servo for three forms of input signal.

621.314.634.011.4 1250

**The Capacitance of Selenium Rectifiers**—R. E. Burgess. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1036–1037; December, 1950.) Comment on paper abstracted in 744 of April (Cooper).

621.316.722 1251

**Stability of Voltage Regulators**—F. E. Bothwell. (*Elec. Eng.*, vol. 69, p. 1090; December, 1950.) Summary of AIEE 1950 Fall General Meeting paper. The presence of nonlinear control elements in many voltage-regulator circuits has prevented a complete mathematical analysis. Theory of Liapounoff, showing that stability of the continuous periodic solutions of a set of nonlinear differential equations is determined by the stability of its linear perturbation equations, is used to solve the problem of stability of electric circuits subjected to small displacements from equilibrium. As an example, characteristic equations are developed for a voltage-regulated $n$ stage dc generator, and the stability boundary is obtained for the case of a step-resistor control element.

621.316.722.1 1252

**Stabilized Power Pack**—R. E. Harvison. (*Electronic Eng.* (London), vol. 22, no. 274, pp. 512–513; December, 1950.) An analysis of the operation of two stabilizing tubes in parallel to provide high current output.

621.316.722.1.076.7 1253

**Precision A.C. Voltage Stabilizers**—G. N. Patchett. (*Electronic Eng.* (London), vol. 22, nos. 271–274, pp. 371–377, 424–428, 470–473, 499–503; September to December, 1950.) Critical study of different types and circuit arrangements providing stabilization to within about 0.01 per cent. *See* also 227 of February.

621.319.45.025:621.314.64 1254

**Rectifier Properties of the System Al-$Al_2O_3$-Electrolyte Subjected to an Alternating Voltage**—A. J. Dekker and W. C. van Geel. (*Philips Res. Rep.*, vol. 5, pp. 303–314; August, 1950. In French.) An account is given of an experiment showing that the static and dynamic current or voltage characteristics of this system are different. A bridge method is described for measuring the dynamic characteristic; the curve exhibits a loop in the region corresponding to forward current. The effect on the loop of frequency and temperature variations is investigated experimentally, and it is concluded that the structure of the oxide coating is not stable, but depends on the magnitude and direction of the applied voltage.

621.316.722 1255

**Voltage Stabilizers [Book Review]**—F. A. Benson. Publishers: Electronic Engineering, London, 1950, 125 pp., 12s. 6d. (*Electronics*, vol. 23, pp. 154, 158; December, 1950.) "A concise monograph . . . this book will be an extremely handy reference for an engineer or technician."

## TELEVISION AND PHOTOTELEGRAPHY

621.397 1256

**G.E.C. Picture-Telegraph Equipment**—(*Electronic Eng.* (London), vol. 22, no. 274, pp. 503–504; December, 1950.) A general description, with diagrams, of the principles of operation of transmitting and receiving equipment, providing the highest possible definition within CCIT limits. At normal speed a 10×8 inch picture is transmitted in 14 minutes; this time can be halved at the expense of picture quality.

621.397.24/.26:621.396.65 1257

**Portable Microwave Television Link**—(*Elect. Commun.* (London), vol. 27, pp. 295–297; December, 1950.) For another account *see* 459 of March.

621.397.331:535.623 1258

**Photoelectric Method of Scanning and Chromatic Selection for Designs in Colour**—M. Lange. (*Toute la Radio*, no. 150, pp. 343–348; November, 1950.) The pattern from which the various colors are to be selected is fixed to a revolving cylinder moving along its own axis, and is scanned by a brightly illuminated spot. An optical system making use of the toothed-wheel principle derives two rays of light, each of which is directed on to a prism spectroscope. Parts of the spectra so obtained are applied to three (panchromatic) photocells, arrangements being made for varying the parts selected. The outputs from the three cells are amplified and mixed in such a way that the resultant voltage is zero for the color to be selected, increases as this color fades, and reaches a threshold value for white; black and any color other than that required produce a voltage above the threshold value. This voltage actuates a galvanometer mirror reflecting a pencil of light of diamond-shaped cross section through an aperture of the same shape. A beam of light of width, depending on the position of the mirror, is directed on to a light-sensitive film carried by the receiving drum. As the drum rotates, a band is recorded on the film. An interposed revolving slotted disk transforms this band into a series of dots. Various refinements affecting the reproduced pattern are described.

621.397.5:535.624 1259

**Subtractive Color TV**—I. Kamen. (*TV Eng.*, vol. 1, pp. 12–13, 28; October, 1950.) The image is produced by passing light from a projection lamp through three dark-trace tubes in succession, the screens being of different materials corresponding to absorption of the three primary colors. A sequential-frame transmission system is used, and the receiver tubes are fed with the appropriate color signals. A greater luminous efficiency is claimed for the system.

621.397.5(494) 1260

**Present State of Television in Switzerland**—(*Bull. schweiz. elektrotech. Ver.*, vol. 40, no. 25, pp. 1001–1003; December 10, 1949. In German.) General review, with discussion of the technical and financial problems involved.

621.397.6:621.396.67 1261

**Empire State Television**—(*Broadcast News*, no. 61, pp. 41–45; September to October, 1950.) A description, with many photographs, of the replacement of the old 67-foot antenna surmounting the Empire State Building by a 217-foot multiple antenna system for stations WCBS TV, WABD, WJZ TV, WPIX, and WNBT.

621.397.6:621.396.67 1262

**Temporary Vision Aerial**—Bolt. (*See* 1076.)

621.397.6.001.4 1263

**Signal Sources for Television Testing**—Thomasson. (*See* 1202.)

621.397.611.2 1264

**Television Camera Tubes**—L. H. Bedford. (*Wireless Eng.*, vol. 28, no. 328, pp. 4–16; January, 1951.) Paper read at the International Television Congress, Milan, 1949. The effect of a high sensitivity tube in a television camera is to make it possible to use a smaller lens aperture for the same object illumination and thus obtain a greater depth of field. Curves showing depth of field against illumination are drawn for present day camera tubes, the minimum possible illumination being governed by the largest practical aperture which can be used with the tube. The minimum illumination is a criterion of sensitivity. A dimensionless form of the criterion, independent of arbitrarily chosen factors, can be found by taking the ratio of this illumination to that required by an ideal tube, the "Quanticon." Other properties of camera tubes discussed and tabulated are color response, resolution, spurious signal generation, signal and noise ratio, contrast, image persistence, geometrical accuracy, stability, and life.

621.397.611.2:778.2 1265

**Flying-Spot Camera, Type TK-3A**—C. R. Monro. (*Broadcast News*, no. 61, pp. 16–21; September to October, 1950.) Equipment is described for the production of video signals from 35 mm slides by the flying-spot method. The control and operating features are explained and typical applications are illustrated by photographs.

621.397.62+621.317.311+536.58:621.38 1266

**Electronics in Britain**—J. H. Jupe. (*N. Z. elect. Jour.*, vol. 23, p. 235; March 25, 1950.) Short descriptions are given of a method for measuring extremely small direct currents or voltages, electronic equipment for temperature, control, and extension units for television.

The Cossor extension unit for television comprises a coupling unit which reduces the modulation from the receiver to 0.5 v, together with the main unit which includes a loudspeaker and cathode ray tube. The unit is designed to give perfect linearity, so that the extension signal is not affected by any nonlinearity existing in the main receiver. The picture may be much larger than that of the main set and the unit can be made to cover more than one type of television system, with negligible increase in cost. For example, units made for use with the British 405 line 50 frames per second system can be adjusted in a few minutes to work with an American 525 line 60 frames per second receiver. *See* also 1216 of 1949 (Zeluff).

621.397.62 1267

**The Simplification of Television Receivers**—W. B. Whalley. (PROC. I.R.E., vol. 38, pp. 1404–1408; December, 1950.) Reprint. *See* 1271 of 1950.

621.397.62 1268

**The Video Output Stage**—E. T. Emms and E. Jones. (*Electronic Eng.* (London), vol. 22, nos. 272 and 273, pp. 408–413 and 454–460; October and November, 1950.) Values of $t$, the signal build-up time for the various elements of the video stage, are derived from the required over-all build-up time $T$ estimated from the build-up time $\tau$ for the transmitted wave; picture definition being classified broadly in terms of $T/\tau$. Different anode compensation circuits are shown and the indicial response (response to unit-step input) is plotted for each circuit. The design of a critically damped cathode compensation circuit is indicated. For equal output and equal $t$ values, the input of such a stage must be 1.4 times that of a first-order shunt-compensated stage; but since the build-up time of the input signal is finite, this is not necessarily a disadvantage. Required characteristics of the video tube are tabulated. Class-A and class-B modes of operation of this tube as an anode bend demodulator are discussed and theoretical efficiency curves are derived. Secondary factors in design are considered and a design procedure is described for each of the first two types of circuit, component values being calculated for circuits using Type EF80 tubes.

621.397.62:621.396.662 1269

**A Variable-Inductance T.V. Tuner**—D. R. DeTar and H. T. Lyman, Jr. (*Electronics*, vol. 23, pp. 102–106; December, 1950.) Description of a tuner covering all vhf television channels without the use of switches or sliding contacts. It comprises a rf amplifier, mixer oscillator, and first detector. The wide frequency range is obtained by the use of only four variable inductors, each of which consists of a specially wound solenoidal coil with fixed core, the tuning being effected by sliding an Al sleeve between coil and core. Performance details are given and comparison is made with conventional designs.

621.397.7 1270

**WOR's TV Studios**—N. F. Smith. (*Broadcast News*, no. 61, pp. 46–73; September to

October, 1950.) A very comprehensive and detailed description of the New York studios and control rooms, including switching arrangements, film-projection equipment, master control system, and monitoring facilities.

## TRANSMISSION

621.316.726.078.3+621.3.018.41(083.74) 1271

**The Evolution of Frequency Control**—C. F. Booth. (*Proc. IEE* (London), Part III, vol. 98, pp. 1–10; January, 1951.) A historical survey is made of the frequency stability of radio transmitters and the frequency tolerances specified by successive international conferences. The theoretical bandwidth needed by a single channel telegraph transmitter at the higher end of the 4- to 30-mc band is a small fraction of the tolerance (±0.003 per cent) assigned to it at the Atlantic City conference. Any improvement in stability will thus increase the number of stations possible. Simple crystal-controlled drive units capable of achieving the present tolerances are described and drive units for synchronized mf transmitters, frequency standards, and quartz clocks capable of much higher stability are discussed.

## TUBES AND THERMIONICS

535.215.1/.2:621.383.2 1272

**Some Properties of Complex Photoelectric Layers**—A. Lallemand and M. Duchesne. (*Z. angew. Math. Phys.*, vol. 1, pp. 195–201; May 15, 1950. In French.) From experimental results on photoelectric layers of the type Ag to $Cs_2O$ to Cs, it is concluded, (a) that such complex layers, although not obeying Richardson's law, have a well defined work function which can be determined from Fowler's theory; (b) that the layers are sensitive to wavelengths in the neighborhood of 15,000 Å, but that the photoelectric emission, though not completely explained by Fowler's theory, is not appreciably greater than that to be expected from the theory; (c) that the observed thermionic emission can be explained on the assumption that the layers consist of a large number of elements whose work function can be determined from Fowler's theory, or of a very small number of elements with a much smaller work function of the order of 0.4 v.

535.215.4:621.383.4 1273

**A Relationship between the Refractive Index and the Infrared Threshold of Sensitivity for Photoconductors**—T. S. Moss. (*Proc. Phys. Soc.* (London), vol. 63, no. 363B, pp. 167–176; March 1, 1950.) "It is shown that for photoconductive compounds the long wavelength "threshold" of sensitivity $\lambda$ should be related to the refractive index $n$ of the photoconductor, and analysis of the available data shows that for the more highly refractive compounds $n^4\lambda$ always has a value close to 77. As a consequence of this relation, an explanation can be given for the change of the threshold wavelength with temperature which is observed for lead sulphide and similar materials. Suggestions are made as to what materials ought to be suitable for photoconductive detectors at longer wavelengths in the infra red, and some confirmation of the ideas is shown by results obtained with cadmium arsenide, the properties of which are briefly described."

537.525.92:537.533.7:621.396.822 1274

**Experimental Verification of Space-Charge and Transit-Time Reduction of Noise in Electron Beams**—C. C. Cutler and C. F. Quate. (*Phys. Rev.*, vol. 80, pp. 875–878; December 1, 1950.) A cavity, resonant at 4.2 kmc, was arranged to be moved axially along a parallel electron beam from a Pierce gun. The observed variation of noise power in the cavity with distance along the beam is in fair agreement with the sinusoidal variation predicted theoretically. The greatest reduction of noise power obtained was 20 db below shot noise. That the minimum noise power is not zero is largely due to residual partition noise.

537.525.92:621.385 1275

**Congruent Space-Charge Flow**—G. B. Walker. (*Proc. Phys. Soc.* (London), vol. 63, no. 372B, pp. 1017–1027; December 1, 1950.) For irrotational flow, propositions are established regarding, (*a*) rectilinear motion, (*b*) motion with constant current density along lines of flow, and (*c*) representation of lines of flow by the level lines of a harmonic function. Two cases of curvilinear flow from a unipotential cathode are discussed and shown to possess important features regarding magnification and transit time.

621.385.029.63/.64 1276

**Interaction between a Travelling Electromagnetic Wave and a Beam of Electrons Moving in a Cylindrical System Perpendicular to Steady Crossed Electric and Magnetic Fields**—R. Warnecke and O. Doehler. (*Compt. Rend. Acad. Sci.* (Paris), vol. 231, pp. 1132–1134; November 20, 1950.) A traveling-wave tube is described in which a steady magnetic field is produced by the passage of a current $I_m$ through the axial element of a coaxial delay line, and a steady electric field by applying a constant voltage $v_0$ between the inner and outer elements of the line. Relatively high gain and high efficiency are claimed to result. Two or more waves are propagated, depending on the operating conditions. For a particular system operating at about 1.5 kmc, with $I_m=500$ $A$ and $v_0=1.2$ kv, calculation gives a useful power of 100 w, efficiency of 40 per cent, and gain of the order of 2.5 db per cm.

621.385.3:537.212 1277

**On the Electric Field in a Single-Grid Radio Valve**—G. B. Walker. (*Proc. IEE* (London), Part III, vol. 98, pp. 57–63; January, 1951.) The problem of calculating the electrostatic field in a planar triode is solved by assuming a system of line charges at the center of each grid wire and hence deriving a potential function which satisfies all the boundary conditions. This function is expressed in series form, but can be readily used to compute field characteristics to any required accuracy. Tables of the more important characteristics are provided.

621.385.3:621.315.592:518.4 1278

**Graphical Analysis of Transistor Characteristics**—Hunter. (*See* 1101.)

621.385.5:537.212 1279

**On the Electric Field in a Multigrid Radio Valve**—G. B. Walker. (*Proc. IEE* (London), Part III, vol. 98, pp. 64–67; January, 1951.) The question of the extent to which field disturbances caused by the individual wires of a grid penetrate an adjacent grid is examined by considering interelectrode capacitance relations. The effect of such disturbances can be measured, using an electrolyte tank, and can also be computed. Only space charge free fields are considered. and the electrode system is assumed either planar or concentric. It is concluded that Dow's simplified treatment (1904 of 1941) may always be used, at least as a first step, in the analysis of a tube containing two or more grids, the field in the region of one particular grid being afterwards investigated by a more rigorous triode analysis.

621.396.615.14.029.63 1280

**Oscillators for Decimetre Waves with Disk-Seal Valves in Grounded-Grid Circuits**—L. Ratheiser. (*Radio Tech., Vienna*, vol. 26, pp. 519–524; November, 1950.) The special features and the use of the disk-seal triode with glass envelope and of the Telefunken metal ceramic type are described. Advantages of the grounded-grid circuit are stressed, and methods of overcoming the low interelectrode capacitance in this case are considered. These are, (a) "disk-edge" feedback, in which the output and input circuits are coupled capacitively to the grid through an insulating layer; (b) feedback windows, which are recesses in the wall of the grid cylinder; these give stability but limit bandwidth; (c) additional external capacitive feedback as in the Philips Type EC55 tube. Drawings of different assemblies are shown and their construction is discussed, particularly the methods used for obtaining good mechanical and electrical contacts.

621.396.615.141.2 1281

**Some Aspects of Split-Anode Magnetron Operation**—Reich, May, Skalnik, and Ungvary. (*See* 1120.)

621.396.615.141.2:537.525.92 1282

**Effects of Space Charge on Frequency Characteristics of Magnetrons**—H. W. Welch, Jr. (Proc. I.R.E., vol. 38, pp. 1434–1449; December, 1950.) "Properties of the magnetron space-charge swarm which affect the propagation of electromagnetic waves are defined in terms of an effective dielectric constant. The space charge is found to be doubly refractive in nature, the velocity of propagation depending on the direction of propagation of the wave, polarization of the wave, and frequency. Effects of synchronism of the rotating space-charge swarm are discussed qualitatively. Experimental results which check parts of the theory are presented. The relationship of the space charge to the circuit is discussed in terms of the nonoscillating and the oscillating magnetron."

621.396.822 1283

**Threshold Signals. [Book Review]**—Lawson and Uhlenbeck. (*See* 1234.)

## MISCELLANEOUS

621.3.012.3 1284

**Reference Sheets**—(*Electronics, Annual Buyers' Guide Issue*, vol. 23, pp. R1-R40; Mid-month June, 1950.) A collection of 21 of the graphical and tabular sheets that have appeared in recent years in *Electronics*, with a complete index of all published since April, 1930. Many of those now included have been revised.

621.396.69+621.38]:061.5(058.7) 1285

**Alphabetical Listings of All Components, Complete Units, Allied Products, Used in Electronic Equipment for All Purposes**—(*See* 1126.)

621.38/.39 1286

**Electronic Engineering Master Index, 1949. [Book Review]**—Publishers: Electronics Research Publishing Co., New York, N. Y., 1950, 296 pp., $17.50. (*Jour. Frank. Inst.*, vol. 250, p. 587; December, 1950.) The third supplement, with coverage increased to nearly 400 periodicals, and so forth. Asterisks indicate Russian articles of which an English translation is available at Brookhaven National Laboratory.

621.396 1287

**Einführung in die Funktechnik (Introduction to Radio Technology). [Book Review]**—F. Benz. Publishers: Springer Verlag, Vienna, 4th edn, 736 pp., 78s. (*Wireless Eng.*, vol. 27, nos. 325/326, p. 273; October and November, 1950.) "First published in 1937; it has been much enlarged and is a very comprehensive and up to date text book of radio engineering."

621.396 1288

**Fortschritte der Funktechnik und ihrer Grenzgebiete, Band 7/8 (Progress in Radio Technology and Related Fields, Vol. 7/8) [Book Review]**—H. Richter. Publishers: Franckh'sche Verlagsbuchhandlung, Stuttgart, 1950, 387 pp., DM 60. (*Arch. elekt. Übertragung*, vol. 4, p. 340; August, 1950.) Includes surveys of the following subjects, contributed by different authors: broadcasting receiver design, radio wave propagation, fm, television, hf and industrial measurements, radio prospecting, cathode ray oscillography, and magnetic-tape recording.

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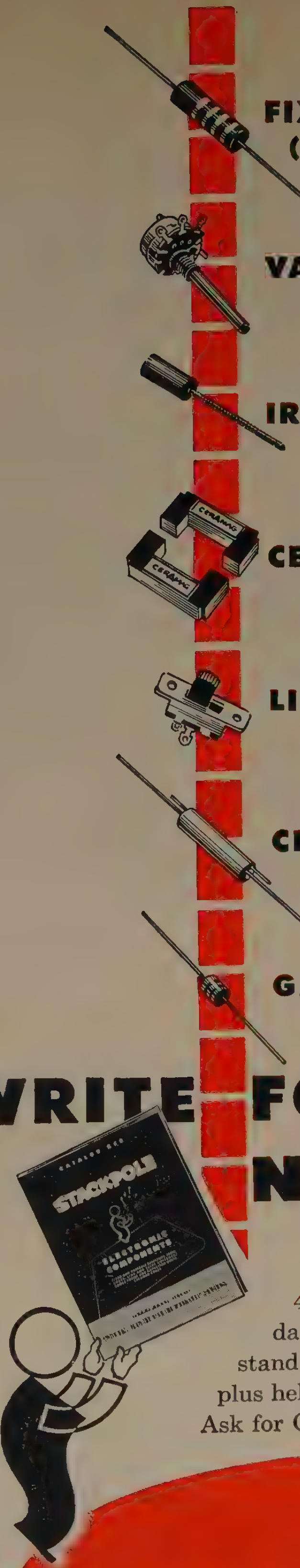

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*(Continued from page 52A)*

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Research electronics engineers with 5-10 years' experience in UHF and pulse techniques or allied fields wanted immediately by research Institute connected with large university in Colorado. Send résumé of education, experience, references, availability and salary requirements to Box 664.

## News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

### Plant Expansion

**Electro-Connector Mfg. Corp.,** has moved to a new location at 190 W. Glenwood Ave., Philadelphia 40, Pa. This move increases their facilities about five times over the old plant.

A major part of this new space has been allotted for greater output of their regular line of tube sockets, plugs, connectors, and tube shields.

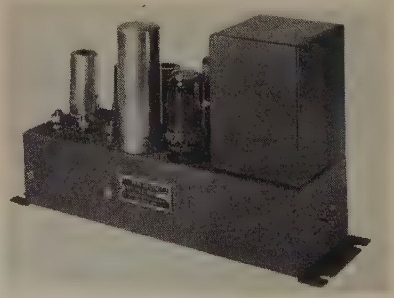
A special products division for production of special aircraft, electronic, and electrical components, and a division for manufacture of a new type of connector were announced.

At the time of the move, the **ELCO Corporation** was formed. The function of this firm, located at the same address, will be to handle sales for Electro-Connector.

### Servo Amplifier

The 410-B, a 60 cps servo amplifier designed to drive a motor with 5-watt output, is announced by **Industrial Control Co.,** 26–02 Fourth St., Long Island City 2, L. I.

It features controls on the gain, carrier phase shift, and damping, with the adjustments independent one from the other. With this range of adjustment, the 410-B can be used in all servo loops in this power class.



Specifications are as follows: maximum gain: 1,000; phase variable from +20 to −140°: internal pickup below 3 millivolts rms; moderate power supply requirements. Companion power supplies and modulators are available.

*(Continued on page 60A)*

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# News—New Products

*(Continued from page 54A)*

## Oscilloscope Calibrator

The A-42F oscilloscope calibrator, a source of continuously variable monitored voltage from 0.0005 volt to 100 volts rms, priced at $24.95 complete with tube and instructions, may be purchased direct from **Tensor Electric Development Co., Inc.**, 343 Classon Ave., Brooklyn 5, N. Y.

When used with any oscilloscope, it enables the operator to make measurements of voltages within this range. It is calibrated directly in peak-to-peak volts and decibels, as well as rms volts.

*(Continued on page 75A)*

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# Positions Wanted By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants, and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The Institute publishes free of charge notices of positions wanted by I.R.E. members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The Institute necessarily reserves the right to decline any announcement without assignment of reason.

### ENGINEER

3 years' experience as Production Supervisor and test B.S.T.E. A.T.I.T., some experience at circuit design, 7 years' analyzing and repair TV and radio receivers. Age 27, married, 1 child Desires either production or design. Midwest. Box 506 W.

### ELECTRONICS ENGINEER

Electronics engineer—physicist. Physics degree from Brooklyn College. 2 years' Army electronics officer, Harvard, M.I.T. radar school. Continuing graduate work. 3 years' design and development work in radar with emphasis on microwave circuitry. Box 516 W.

### ELECTRICAL ENGINEER

M.S. in E.E., University of Illinois. 1 year as industrial development electrical engineer. 2 years' full time university teaching. Single. Age 28. Desires a position as development engineer involving automatic control work. Box 517 W.

### ELECTRICAL ENGINEER

B.S. and M.S. Cal. Tech. 3 years' Signal Corps, 1 year own business, 3 years' teaching. Southern California area. Possibly invest in business. Box 518 W.

### ENGINEER

B.E.E. electronic option, University of Kentucky, June 1950. Age 31, married; U.S. Civil Service training, field radio and frequency modulation; 3 years' Army Signal Corps, installation and repair. Desires position in design and development. Résumé on request. Box 519 W.

### ENGINEER

M.S. in E.E. University of Michigan, June 1951. Communications field major. 1½ years' teaching Navy electronics, research in upper atmosphere rocket program. Married, aged 24. California area. Box 520 W.

### TELEVISION ENGINEER

B.S.T.E. A.T.I.T. 1st class radio-phone license. 1 year experience TV transmitter and studio equipment. 2 years' experience TV service supervisor. Age 26, married, 1 child. Desires either TV design or position in TV station. Midwest. Box 526 W.

*(Continued on page 62A)*

**ENGINEERING OPPORTUNITIES IN Westinghouse**

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Must have at least one year's experience.

For work on airborne radar, shipborne radar, radio communications equip., microwave relay, or micro-wave communications.

Good pay, excellent working conditions; advancement on individual merit; location Baltimore.

Send resume of experience and education to: Manager of Industrial Relations, Westinghouse Electric Corp., 2519 Wilkens Ave., Baltimore 3, Maryland.

**NATIONAL UNION RESEARCH DIVISION**

The Research Laboratories of one of the nation's larger electron tube manufacturers have vacancies for electronic engineers and engineering physicists qualified in the following fields:

**Vacuum tube development**
**Electronic circuit design**
**Electronic equipment**

This organization can offer excellent prospects for security and personal advancement due to our continued growth. Our location is in the New York metropolitan area.

Whether you have a background of electron tube or circuit design, or are a recent graduate and interested in our field, we would like to hear from you. Send your résumé to:

*Personnel Department*
NATIONAL UNION
RESEARCH DIVISION
350 Scotland Rd. Orange, N.J.

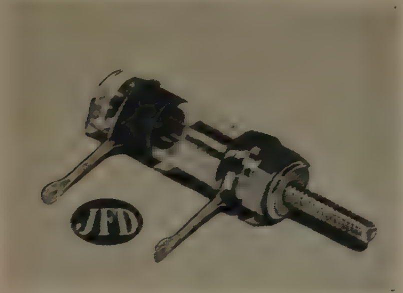
# News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 60A)*

## Piston Type Variable Trimmer Capacitors

Three new variable trimmer capacitors of the piston type are being produced by the **JFD Manufacturing Co., Inc.,** 6101 16th Ave., Brooklyn 4, N. Y.



All have a single-unit movable electrode, and adjustment screw made of an alloy with a low temperature coefficient. A silver band fused to the exterior of the quartz tube serves as a stationary electrode.

The VC 3 has a maximum of 0.3 $\mu\mu f$ at minimum setting, and a minimum 3 $\mu\mu f$ at maximum when measured at 0.1 to 1 mc. The VC 5 has a maximum of 0.5 $\mu\mu f$ at minimum setting, and a minimum of 5 $\mu\mu f$ at maximum. The VC 11 has a maximum of 1.5 $\mu\mu f$ at minimum setting and a minimum of 11 $\mu\mu f$ at maximum setting.

Temperature coefficient approximately zero; "Q" of over 1,000 at 1 mc. operating temperature −55° to +100°C. Will mount on panels of ⅛ inch or less in thickness, and extends 15/16 inch in back of panel.

## Inductrol Power Packs

A new "midget" load center unit substation, designed for low-voltage, regulated ac lighting and power service in factories and laboratories has been announced by the **General Electric Co., Unit Equipment Div.,** Schenectady 5, N. Y.

Called an Inductrol Power Pack, the unit incorporates in one steel housing an air circuit breaker, a dry-type transformer, and an air-cooled induction regulator.

*(Continued on page 80A)*

PRECISION ELECTRONIC SYSTEMS

RADIO
- Direction Finders Electronic Goniometer
- Signal Generators Crystal Control 1 kc to 150 mc.
- Spectrum Generator Marker Pulses .5 to 50mc.

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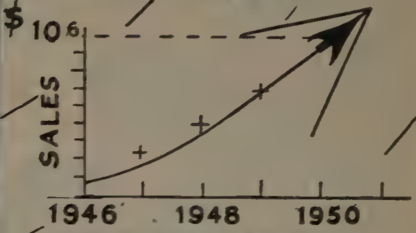

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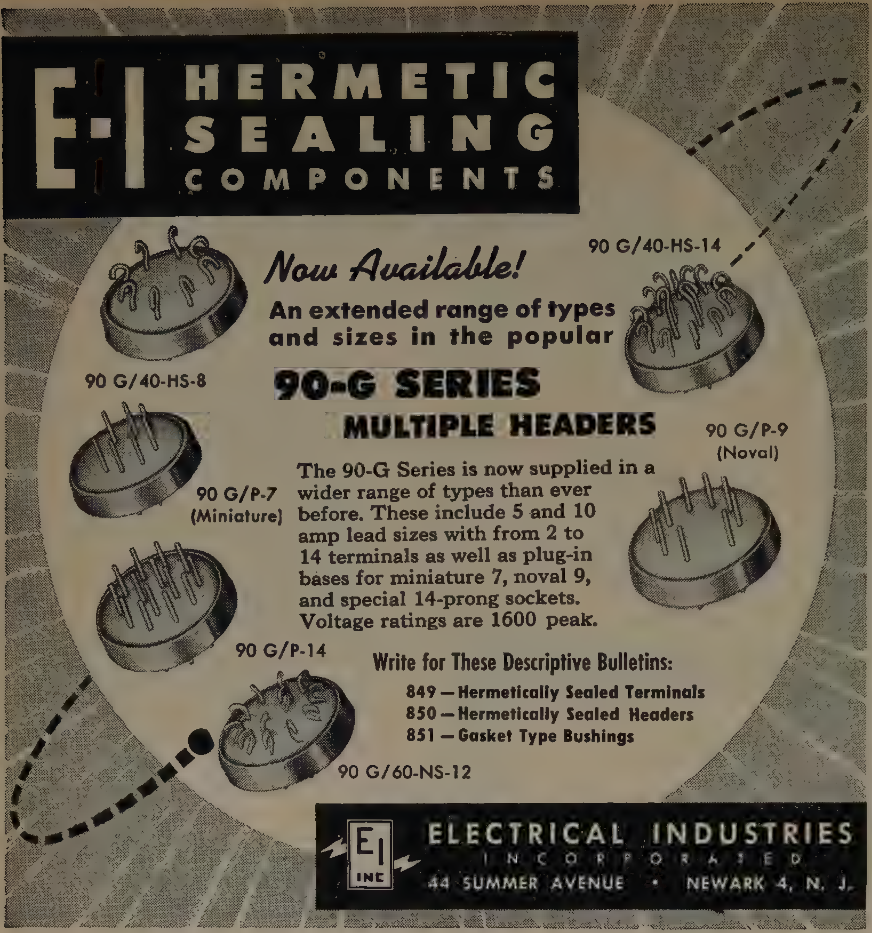
E-I HERMETIC SEALING COMPONENTS

Now Available!

An extended range of types and sizes in the popular

90-G SERIES MULTIPLE HEADERS

The 90-G Series is now supplied in a wider range of types than ever before. These include 5 and 10 amp lead sizes with from 2 to 14 terminals as well as plug-in bases for miniature 7, noval 9, and special 14-prong sockets. Voltage ratings are 1600 peak.

Write for These Descriptive Bulletins:

849—Hermetically Sealed Terminals
850—Hermetically Sealed Headers
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ELECTRICAL INDUSTRIES INCORPORATED
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90 G/P-14
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*aid precision control and instrumentation*

If you are being called upon to design and build precision electrical and electronic apparatus for tough field service, be sure to write for a copy of the Ney book, "Precious Metals for Sliding Contacts and Non-Corrosive Wear Resisting Parts." New experience and wealth of successful application data in this specialized field may be the solution to one of your important problems. The book includes comprehensive technical data on Ney precious metal alloys together with photographs and dimension drawings of a wide variety of contacts, brushes and other components now standardized for production.

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*(Continued from page 73A)*

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*(Continued on page 78A)*

THE RIGHT EQUIPMENT

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20,000 OHMS/VOLT

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KIT OR WIRED

This laboratory-precision instrument is a basic tool necessary for up-to-the-minute engineering. It incorporates 31 different ranges . . . offering the greatest number of ranges and versatility. The "open" design permits complete accessibility to components.

The 4½" meter has a 50 ua D'Arsonval movement for dependable accuracy. All resistors have 1% or better accuracy.

Ruggedly constructed . . . handsome, durable, high-impact Bakelite case and panel . . . imbedded figures and symbols for extra-long wear and readability.

RANGES:

*DC Voltage:* 0-2.5, 10, 50, 250, 1000, 5000, at 20,000 ohms/volt

*AC Voltage:* same ranges at 1000 ohms/volt

*Output Voltage:* same as AC ranges, with internal isolation condenser

*Decibels:* —12 db to —55 db in 5 ranges

*DC Resistance:* R x 1, R x 100, R x 10,000

*DC Current:* 0-100ua, 10ma, 100ma, 500ma, 10A (250 millivolts)

MODEL 555K
Kit . . . complete with batteries $29.95

MODEL 555
Factory wired . . . complete with batteries $34.95

DIMENSIONS:
6¾ x 5¼ 3"

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(Continued from page 78A)

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Radio Engineers

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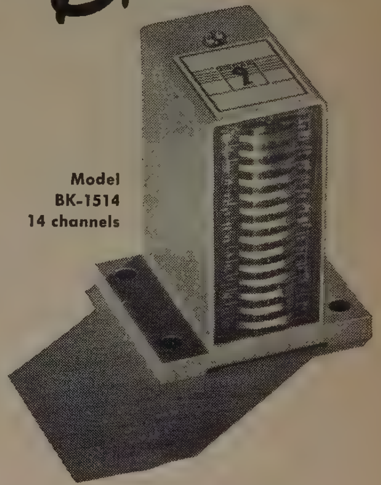
MULTICHANNEL RECORDING COMPONENTS

BY *Brush*



Model
BK-1514
14 channels

- All gaps in exact alignment
- Cast in a block of synthetic resin
- Magnetic shielding between channels
- Nonmicrophonic
- Adaptable to many widely different applications including sound and pulse recording
- Channels available up to 14

*Brush also introduces:*

Magnetic Recording Rubber (Synthetic)
Watch for new announcements.

*For information about these and other Brush Magnetic Recording Components in stock or for special purposes, write*

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SELENIUM RECTIFIERS

Performance

is no accident...

After all, there can be but one standard of quality for Selenium Rectifiers...the highest. No matter how high the quality of the other component parts of the unit requiring a Selenium Rectifier...or how skillfully that unit is designed and engineered ...if the Rectifier fails, the unit is useless.

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TYPE V-HF SERIES
5 MILLIAMPERES DC

Circuit-Half-Wave. In 9/16" OD Phenolic Tube with ferrule at each end for insertion in Fuse Clips. Overall length varies to 9" depending on the DC output voltage rating.

PARTIAL LIST OF TYPE V-HF SERIES RECTIFIERS AVAILABLE

| DC Output Voltage | Rectifier Part No. | DC Output Voltage | Rectifier Part No. |
|---|---|---|---|
| 40 | V2HF | 800 | V40HF |
| 80 | V4HF | 1500 | V75HF |
| 200 | V10HF | 2500 | V125HF |
| 600 | V30HF | 4000 | V200HF |

TYPE Y-HP SERIES
11 MILLIAMPERES DC

Circuit-Half-Wave. In 9/16" phenolic tube with pigtail leads. The overall length of rectifiers in this series varies up to 9", depending on the DC output voltage rating.

PARTIAL LIST OF Y-HP SERIES RECTIFIERS AVAILABLE

| DC Output Voltage | Rectifier Part No. | DC Output Voltage | Rectifier Part No. |
|---|---|---|---|
| 20 | Y1HP | 1000 | Y50HP |
| 60 | Y3HP | 2000 | Y100HP |
| 100 | Y5HP | 3000 | Y150HP |
| 400 | Y20HP | 4000 | Y200HP |

*We manufacture over 500 types of Rectifiers for Special Requirements. Write for further data and literature.*

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# News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 75A)*

The new pack, available in either single- or three-phase ratings, has a capacity ranging from 15- to 100-kva, with incoming circuit rated 480 or 600 volts, 60 cps, and a regulated output at 120/240 or 208Y/120 volts.

In addition to lighting, the Inductrol Power Pack can be used to regulate the power supplied to resistance heating and infrared heating equipment, electronic apparatus, precision instruments and control circuits, and commercial electric cooking epuipment.

## 3KW Multichannel Transmitter

High-power simultaneous multichannel operation is provided by the Model T3M Transmitter, designed by **Press Wireless Mfg. Co., Inc.**, Hicksville, L. I., N. Y. In both telephone and telegraph, four or more transmitting channels are instantly selectable and simultaneous operation of combinations of services can be established by local or remote dialing.

The T3M comprises three basic units, the rf unit, the amplifier-modulator unit, and the power supply unit. One rf unit, and a power supply unit compose a complete radiotelegraph transmitter. The addition of an amplifier-modulator unit provides for radiotelephone operation. Each radio-frequency unit is complemented with a frequency shift keyer adapter for simple connection of an F-S Keyer. The amplifier-modulator is capable of 100 per cent modulation of the full 3-kw output of any rf unit. The power supply unit will simultaneously power any of the following equipment combinations: four rf units operating in A-1 (telegraph); three rf units operating in F-1 (Frequency Shift); by adding an additional amplifier-modulator unit, two rf units can be operated in A-3. Frequency Range—2 to 20 mc in three ranges. Power output—3 kw, A-1, A-3, or F-S. Frequency control—temperature regulated quartz crystals. Frequency shift operation—built-in frequency shift keyer adapter. Modulation capability—100 per cent, high level. Carrier noise—down 40 db from 100 per cent modulation. Keying—electronic keying.

*(Continued on page 82A)*

Where the Requirements are Extreme...

Use
SILVER
GRAPHALLOY

For extraordinary electrical performance

GRAPHALLOY
TRADE MARK REG. U. S. PAT. OFF.

THE SUPREME BRUSH AND CONTACT MATERIAL

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- minimum wear
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Accumulated design experience counts—call on us!

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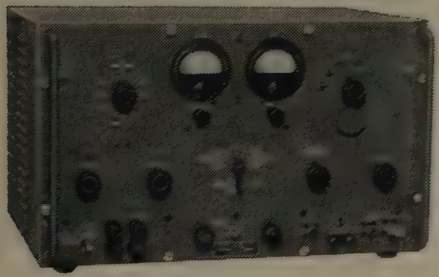
## CHECK OUT

## WITH SIGNAL GENERATORS

by

**AIRCRAFT RADIO Corporation**



**TYPE H-14** 108-132 MEGACYCLES

Standard signal source for complete testing of VHF airborne omnirange and localizer receivers in aircraft or on the bench is ARC's Type H-14 Signal Generator. It checks up to 24 omni courses, omni course sensitivity, to-from and flag-alarm operation, left-center-right on 90/150 cycle and phase-localizers, and all necessary quantitative bench tests. Permits quick, accurate, check-out of aircraft just before take-off. For ramp checks RF output 1 volt into 52 ohm line; for bench checks, 0-10,000 microvolts. AF output available for bench maintenance and trouble shooting.

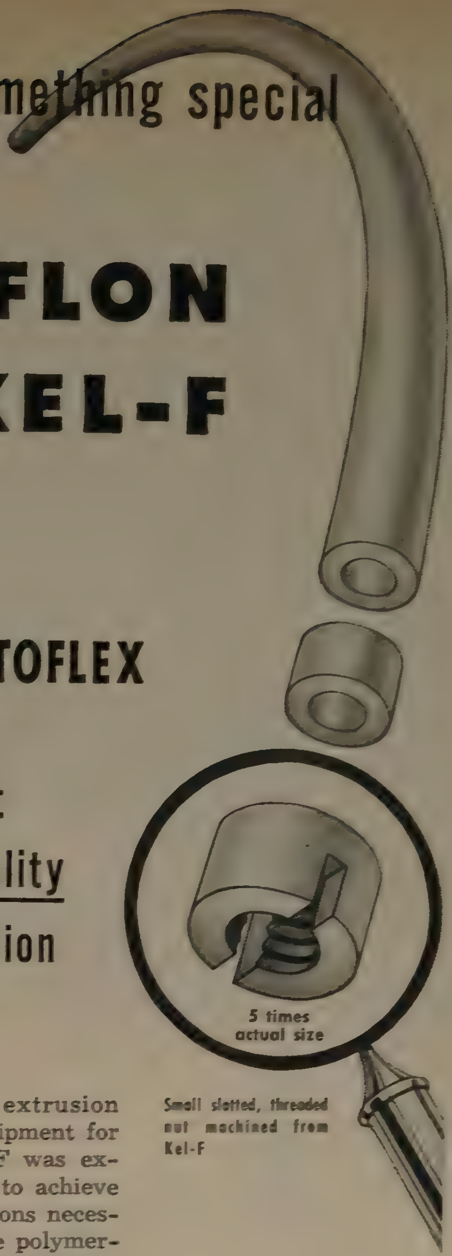
**Price $885.00 net, f.o.b. Boonton, N. J.**

**Type H-12** VHF Signal Generator 900—2100 mc—source of cw or pulse amplitude-modulated RF. Power level 0 to −120 dbm. Internal pulse circuits with controls for width, delay, and rate, and provision for external pulsing. Frequency calibration better than 1%. Built to Navy specs for research, production testing. Equal to Military TS-419/U.

**Price: $1,950.00 net f.o.b. Boonton, N. J.**

**AIRCRAFT RADIO CORPORATION**
Boonton New Jersey
*Dependable Electronic Equipment Since 1928*

# Need something special in TEFLON or KEL-F



Small slotted, threaded nut machined from Kel-F

## Let RESISTOFLEX make it and get top quality production

At Resistoflex, extrusion and molding equipment for Teflon and Kel-F was expressly designed to achieve the exact conditions necessary for complete polymerization without degradation.

This assures you utmost stability and inertness in these fluorocarbon resins for high frequency insulation over a wide range of temperatures, and under the severest corrosive conditions.

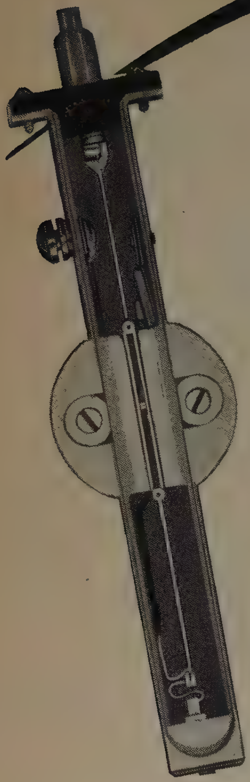
In addition, Resistoflex offers the ultimate tensile strength and "plastic memory" in these thermoplastics. They're free from internal strain which means better machinability and longer service life.

Rigid control over processing conditions combines *uniformity* of their outstanding properties with *dimensional* uniformity to give maximum production schedules and lower fabrication costs.

Write us for a quotation on rods, tubes, sheets, molded and machined parts produced from Teflon and Kel-F under our "FLUOROFLEX" trade mark.

*For out of the ordinary engineering with synthetics*

**RESISTOFLEX CORPORATION**
Belleville 9, N. J.



Sensitive Rugged Dependable

HEILAND GALVANOMETERS

of high current and voltage sensitivity are the heart of Heiland recorders.

Heiland galvanometers are available in natural frequencies ranging from 32 cycles per second to 4700 cycles per second. All galvanometers are electromagnetically damped with exception of those having natural frequencies above 450 cycles per second; these are fluid damped. Permanent magnet assemblies are available for mounting 2, 6, and 12 Heiland galvanometers.

dependable instruments

Heiland DENVER

For further information on Heiland Galvanometers, Multi-Channel Oscillograph Recorders or Associate equipment, write or wire . . .

HEILAND RESEARCH CORPORATION 130 East 5th St. DENVER, COLO.

Wayne Kerr of England

V.H.F. BRIDGES . . .

Transformer Ratio-Arm principle covers frequency range of

15Kc. to 250Mc.

Wayne Kerr Laboratories have recently developed a range of H.F. and V.H.F. Transformer Ratio-Arm Bridges which are as stable in operation as the normal type of low frequency bridge. This has been achieved with a design which gives an extremely low impedance across the bridge terminals and between terminals and ground. Consequently, freedom from errors due to parasitic reactances permits a much greater accuracy of measurement. Up to 120 Mc/s balanced or unbalanced measurements can be made with equal facility.

*Demonstrations in New York by the U.S. Distributors:—*

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THE WAYNE KERR LABORATORIES LTD. NEW MALDEN SURREY ENGLAND

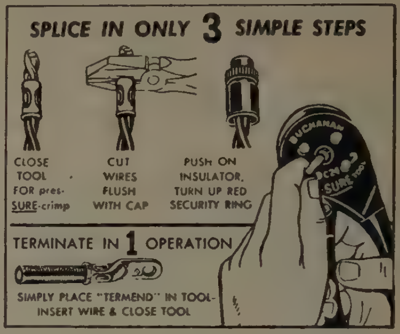
## News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 80A)*

### Pressure Connectors

Splice caps for "pigtail" splicing of electrical wires are now available in an improved open-end construction from **Buchanan Electrical Products Corp.,** 1290 Central Ave., Hillside, N. J.


SPLICE IN ONLY 3 SIMPLE STEPS
CLOSE TOOL FOR pres-SURE-crimp
CUT WIRES FLUSH WITH CAP
PUSH ON INSULATOR, TURN UP RED SECURITY RING
BUCHANAN
SURE TOOL
TERMINATE IN 1 OPERATION
SIMPLY PLACE "TERMEND" IN TOOL-INSERT WIRE & CLOSE TOOL

These open-end splice caps insure that wire insulation is always flush with the cap for maximum circuit protection and that wires are inserted to the full depth of the cap for maximum joint efficiency.

Only two sizes are required for most frequently used combinations of two or more wires ranging from two #18 to three #8.

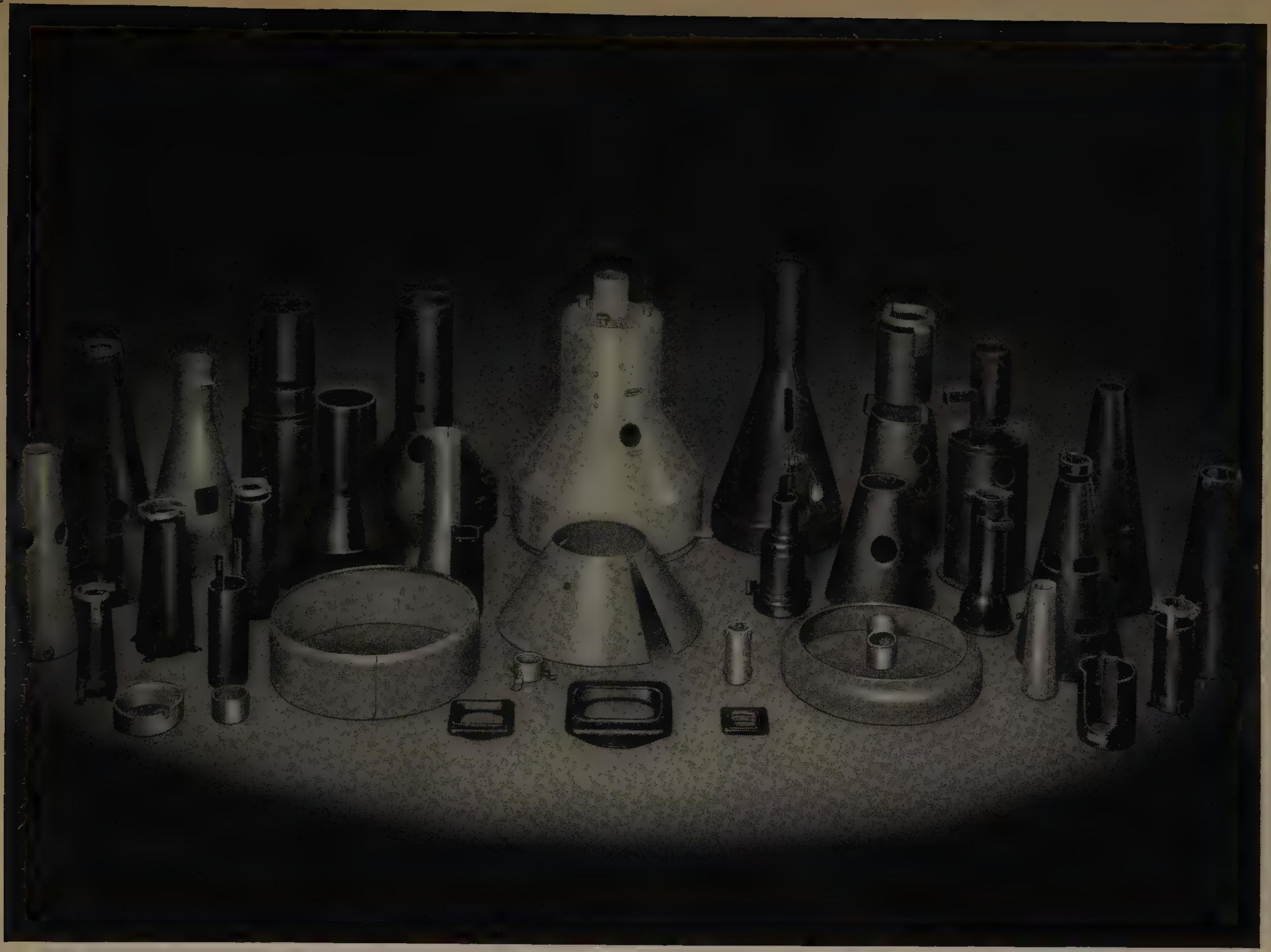
The hand-operated "pres-SURE-tool" which installs both sizes of Splice Caps also installs the manufacturer's Termend lugs on all wire sizes from #16 to #8. This tool has a four-way "pres-SURE-crimping" action which insures connections of maximum electrical and mechanical efficiency and is equally effective on solid or stranded or on combinations of solid and stranded wires.

### Silvered Mica Capacitor

**Cornell-Dubilier Electric Corp.,** South Plainfield, N. J., announces a new Silver Mike, a Type 22R high-stability silvered mica capacitor in less than postage stamp size—$\frac{1}{2} \times \frac{5}{16} \times \frac{3}{16}$ inch. These capacitors are designed to be used in high-$Q$ electronic circuits, where frequency stability and minimum loss must be maintained.

Each unit is wax-sealed against humidity, and the ruby mica dielectric is uniformly silvered on both sides, with contact foils firmly connected to the leads. Type 22R is made to withstand an operating temperature of 100° C. and is tested for capacity, tolerance and at twice the rated voltage. Ratings are: 300-volt dc working from 5 $\mu\mu$f to 510 $\mu\mu$f, and 500-volt dc working from 5 $\mu\mu$f to 420 $\mu\mu$f. The standard capacity tolerance is ±5 per cent. It also is available in ±20 per cent, ±10, ±3, and ±2 per cent, or ±1 $\mu\mu$f (whichever is closer).

*(Continued on page 86A)*



## *Designed for Application*

## **Mu Metal Shields**

The James Millen Mfg. Co. Inc. has for many years specialized in the production of magnetic metal cathode ray tube shields for the entire electronics industry, supplying magnetic metal shields to manufacturing companies, laboratories and research organizations. Stock shields are immediately available for all of the more popular sizes and types of cathode ray tubes as well as bezels for 2″, 3″ and 5″ size tubes.

Many production problems, however, make desirable special shields designed in conjunction with the specialized requirement of the basic apparatus. Herewith, are illustrated a number of such custom built shields. Our custom design and fabrication department is at the service of our customers for the development and manufacture of magnetic metal shields of either nicoloi or mumetal for such specialized applications.

**JAMES MILLEN MFG. CO., INC.**

MAIN OFFICE AND FACTORY
MALDEN, MASSACHUSETTS, U.S.A.

Miniature Transformers and Coils to your Specifications!

Here is a line of sub-miniature transformers developed especially to meet the requirements of today's electronic gear, where space and weight is at a premium.

We can produce to your specifications

Hermetically Sealed Transformers, Radar Deflection Yokes and Coil Components, Television Flyback Transformers and Television Focus Coils.

Custom-made parts to JAN and MIL specifications.

Submit blueprints and quantities for estimates.

**TETRAD CO., INC.**

4921 EXPOSITION BLVD.
LOS ANGELES 16, CALIF.

---

S·S·White MOLDED RESISTORS

The "All-Weather" Resistors

ARE USED IN THIS ULTRA SENSITIVE ELECTRONIC PHOTOMETER

In this instrument—designed for measurement of very low light values—S.S.White Resistors serve as the grid resistance in the all-important high-gain D.C. amplifier circuit. The manufacturer, Photovolt Corp., New York, N.Y., reports that the resistors "work very satisfactorily"—which checks with the experience of the many other electronic equipment manufacturers who use S.S.White resistors.

WRITE FOR BULLETIN 4906

It gives essential data about S.S.White Resistors, including construction, characteristics, dimensions, etc. Copy with price list on request.

*Photo courtesy of Photovolt Corp., New York, N.Y.*

S.S.WHITE RESISTORS are of particular interest to all who need resistors with inherent *low noise level* and *good stability* in all climates.

HIGH VALUE RANGE
10 to 10,000,000 MEGOHMS
STANDARD RANGE
1000 OHMS to 9 MEGOHMS

THE S.S.White INDUSTRIAL DIVISION
DENTAL MFG.CO.
Dept. G-R, 10 E. 40th St.
NEW YORK 16, N.Y.

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# News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 82A)

## Gear Trains

Sealed, self-oiling gear trains in a wide variety of standard speedchanging ratios for output shaft torque loads up to 0.375 pound inches are now available from **Telechron Inc.**, Ashland, Mass., makers of electric clocks, timers, timing motors, and instrument movements.

The standard ratios range from 7,200:1 to 25:1. Other ratios as high as 216,000 to 1 may be obtained, depending on the load involved.

In addition to their sealed, self-oiling design, these compact gear trains utilize hobbed gears only—not stamped gears—for accurate, quiet operation and long life in applications calling for the transmission of small amounts of power.

(Continued on page 88A)

---

Over 85% of the torque wrenches used in industry are

STURTEVANT TORQUE WRENCHES

Read by Sight, Sound or Feel.

- Permanently Accurate
- Practically Indestructible
- Faster—Easier to use
- Automatic Release
- All Capacities

in inch ounces . . . inch pounds . . . foot pounds (All Sizes from 0-6000 ft. lbs.)

STURTEVANT TORQUE MANUAL

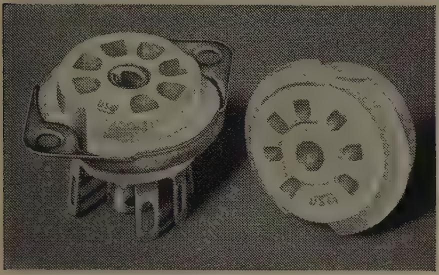
Every manufacturer, design and production man should have this valuable data. Sent upon request.

P.A. STURTEVANT CO.
ADDISON QUALITY ILLINOIS

**first again with**

# TEFLON

7-pin Miniature Tube Sockets, ideal wherever high or low ambient temperatures or frequency stability are problems.



## CHEMELEC

### HIGH PERFORMANCE

### TUBE SOCKETS

- Surface resistivity 3.5 x $10^{13}$ ohms.
- Loss factor less than 0.0005.
- Dielectric constant 2.0 (60 cycles to 30,000 megacycles).
- Serviceable at −150°F to +575°F with negligible change in critical electrical characteristics.
- Won't carbonize under arcing.
- 0.0% water absorption (ASTM Test).
- Unaffected by extreme humidity, corrosive atmosphere or fungus.
- Non flammable.
- Chemically inert—unaffected by oils, organic or inorganic acid, alkali or solvent.
- Allocation limits supply to projects with National Production Authority sanction.
- Write for catalog No. SO-428.

**TEFLON PRODUCTS DIVISION**

**UNITED STATES GASKET CO.**

**611 N. 10th Street, Camden, New Jersey**

---

Standard

# RADIO INTERFERENCE and FIELD INTENSITY Measuring Equipment

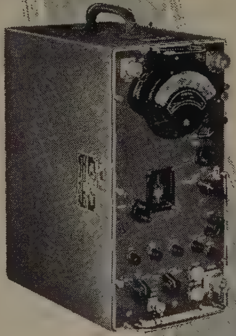
**Complete Frequency Coverage - 14kc to 1000mc!**

**VLF!**

**NM - 10A**

**14kc to 250kc**

Commercial Equivalent of AN/URM-6.
Very low frequencies.



**HF!**

**NM - 20A 150kc to 25mc**

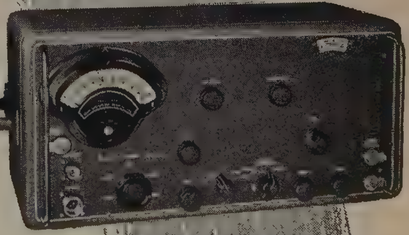
Commercial Equivalent of AN/PRM-1. Self-contained batteries. A.C. supply optional. Includes standard broadcast band, radio range, WWV, and communications frequencies.

**VHF!**

**NMA - 5**

**15mc to 400mc**

Commercial Equivalent of TS-587/U.
Frequency range includes FM and TV Bands.



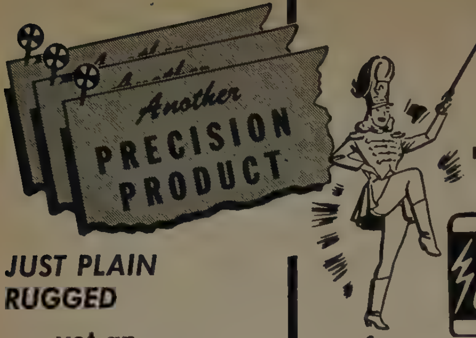
**UHF!**

**375mc to 1000mc NM - 50A**

Commercial Equivalent of AN/URM-17.
Frequency range includes Citizens Band and UHF color TV Band.

**These instruments comply with test equipment requirements of such radio interference specifications as JAN-I-225, ASA C63.2, 16E4(SHIPS), AN-I-24a, AN-I-42, AN-I-27a, AN-I-40 and others.**

**STODDART AIRCRAFT RADIO CO.**
6644-C SANTA MONICA BLVD., HOLLYWOOD 38, CALIFORNIA
Hillside 9294



FOLLOW THE PARADE TO THE *New* Burlington *Hermetically Sealed* INSTRUMENT

JUST PLAIN RUGGED

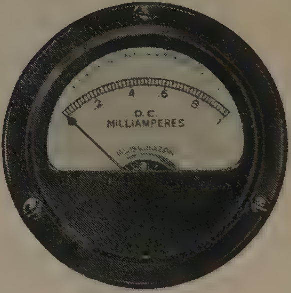
. . . *yet an instrument of highest QUALITY*

. . . *PRECISION*

. . . *ACCURACY*

The Burlington "Hermetically Sealed" *Instrument* was designed and is manufactured to conform to JAN specifications for sealed *instruments.*

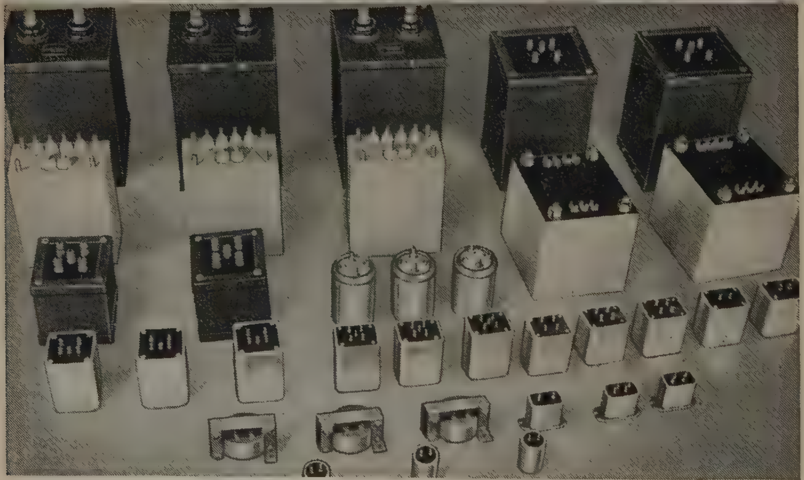
- Steel case with heavy copper-cadmium plate and black finish.
- Excellent shielding due to case material and construction.
- Double strength clear glass.
- Black satin onodized aluminum bezel.
- Glass to metal seal under controlled humidity and temperature conditions.
- D'Arsonval permanent magnet type movement for DC applications.
- Designed to enhance panel appearance.
- Available in 1½" square, 2½" and 3½" round case types.
- Guaranteed for one year against workmanship and materials.



*Best Buy Burlington*

BURLINGTON INSTRUMENT COMPANY
DEPT. I-61, BURLINGTON, IOWA

PERFORMANCE TESTED TRANSFORMERS DESIGNED FOR YOUR NEEDS



QUALITY PRODUCTS PRODUCED IN QUANTITY

ACME ELECTRIC CORPORATION
446 WATER STREET CUBA, N. Y.

Acme ACME Electric TRANSFORMERS

## News—New Products

*(Continued from page 86A)*

### Selenium Rectifier Handbook

The Rectifier Div., Sarkes Tarzian, Inc., 415 N. College Ave., Bloomington, Ind., has recently published a 64-page selenium rectifier handbook containing information on power conversion and suggested applications of selenium rectifiers.

SELENIUM RECTIFIER
handbook
Sarkes
Tarzian
RECTIFIER DIVISION, BLOOMINGTON, INDIANA

In addition to complete information on selenium rectifiers which have found wide use in radio and television receivers, this handbook contains information on power

*(Continued on page 89A)*

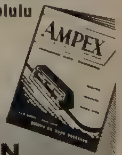
1st.. IN HIGH QUALITY RECORDING

STANDARD OF THE GREAT RADIO SHOWS

AMPEX

MAGNETIC TAPE RECORDERS

SOLD BY . . .
BING CROSBY ENTERPRISES, Inc., Hollywood
AUDIO-VIDEO PRODUCTS CORP., New York City
GRAYBAR ELECTRIC CO., Inc., Principal Cities
THE HAWAIIAN NETWORK, Ltd., Honolulu
RADIO PARTS, Inc., New Orleans
Export: WESTREX CORPORATION, New York City



AX 33

SEND FOR BOOKLET

AMPEX
ELECTRIC CORPORATION
San Carlos • California

# News—New Products

(Continued from page 88A)

rectifiers for high-current applications and high-voltage enclosed rectifiers for low-current electronic applications. Price, 25 cents.

## Power Amplifier

The Type 1233-A power amplifier, a product of the **General Radio Co.**, 275 Massachusetts Ave., Cambridge 39, Mass., is a wide-frequency-range aperiodic amplifier providing a maximum of 15 watts output with less than 0.2 volt input.

In addition to its use as a general laboratory or testing amplifier it has specific applications in the driving of supersonic generators, the exciting of broadcast antennas for measurements with deflection-type instruments, and for use as an oscilloscope deflection amplifier. When used with an antenna and tuned input it can drive a Type 1931-A modulation monitor to monitor remote transmitters.

Three output combinations are available with a maximum frequency range from 20 cps to 3 mc. Output transformers furnish low-impedance output up to 1.5 mc. Noise and distortion are low on all ranges. Readings of output voltage are given on a 3-range panel vacuum-tube voltmeter.

(Continued on page 90A)

Langevin
AMPLIFIERS
For broadcast, public address, recording, and music services — custom designs for special applications.

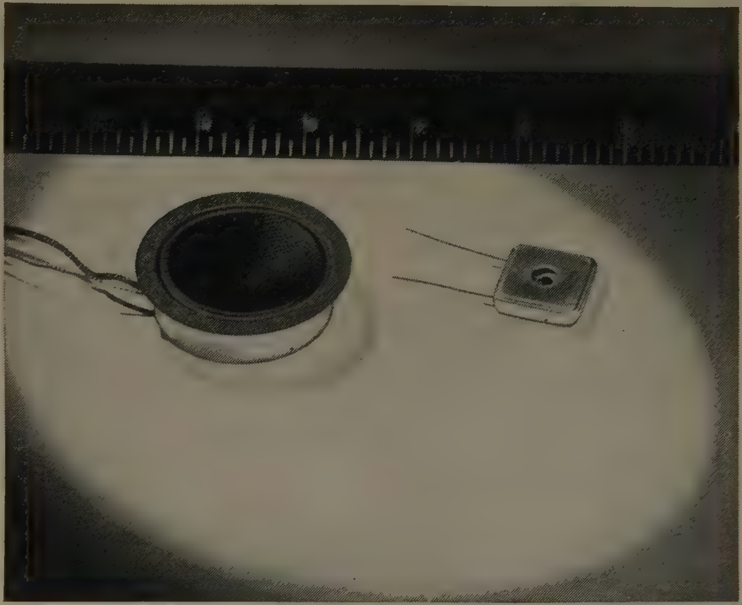
Langevin
TRANSFORMERS
Open-core, encased, hermetically sealed, high-temperature. Built to your own or MIL-T-27 specifications.

Langevin
ENGINEERING
Available for the development and manufacture of special electronic devices.

For detailed information on our products and services, write to

Langevin
MANUFACTURING CORPORATION
37 W. 65th St., New York 23, N. Y.

hearing aids have shrunk to half their size



A Brush DEVELOPMENT

● Brush has been the leader in cutting the size of hearing-aid microphones. Now hearing aids are small, almost unobservable and, most important of all, do their job better. Deaf people no longer have to wear bulky, unsightly and hard-to-operate hearing aids.

The Brush Development Company for more than 21 years has studied the intricate, though powerful, piezoelectric effects of crystals. The principle of this smaller, more powerful microphone, as pioneered and developed by Brush, is that a tiny, jewel-like crystal converts sound waves into powerful electric signals. This is an example of continued engineering effort by Brush to make available to industry the many benefits and uses of piezoelectric materials, whether in hearing aids, measuring instruments or recording devices.

Manufacturers of

ACOUSTICAL EQUIPMENT
Microphones — General Purpose and Specialized
Hearing Aid Microphones
Earphones — Of Many Types

MAGNETIC RECORDING DEVICES
Tape
Plated Wire, Disc and Drum
Multichannel Recorders
Recording and Erase Heads
Memory Storage Units
Computer Components

RESEARCH AND INDUSTRIAL INSTRUMENTS
Universal Strain Analyzers
Surface Analyzers
Multichannel Direct Writing Oscillographs
AC, DC and Carrier-type Amplifiers
Uniformity Analyzers
Transient Recorders

PIEZOELECTRIC CRYSTALS

ULTRASONIC EQUIPMENT
Generators and Analyzers
Laboratory Equipment

OUR BUSINESS IS THE FUTURE
THE BRUSH DEVELOPMENT CO.
3405 PERKINS AVENUE • CLEVELAND 14, OHIO

IT'S TIME TO

PLAY BALL

WITH YOUR DISTRIBUTOR!

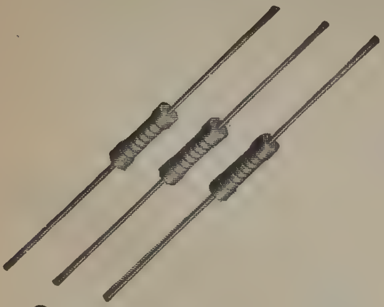
Your distributor is doing everything in his power to give you service today. A large number of the parts and components he normally carries in stock are on the "critical" list because of the vital requirements of the Armed Forces and Defense Industries. As a result he is working harder, longer and going to far greater lengths than ever before to keep your needs supplied.

Naturally, there will be shortages and delays, but don't blame your distributor—play ball with him and he'll do his best for you. Whatever your needs are for electronic components or sheet metal—place your order promptly with your distributor so that he in turn can order from us and get delivery as soon as possible. Bud makes the widest variety of products for your requirements—very often by asking your distributor you may be able to find an excellent substitute for the part you originally had in mind. Work with your distributor, and he'll work for you!

BUD

BUD RADIO, Inc.
2110 EAST 55th STREET CLEVELAND 3, OHIO

CONTINENTAL NOBLELOY
METAL FILM
NF RESISTOR



- Low Temperature Coefficient
- Range; 1 Ohm to 1 Megohm
- Axial Leads; ½ Watt
- Small Size ½" long x .150" Dia.
- Tolerance; + — 1% & 5%

Continental type NF "Nobleloy" resistors were designed to meet the needs of miniature, stable, precision resistors in critical applications.

Write for further details.

CONTINENTAL CARBON INC. CLEVELAND 11, OHIO

## News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 89A)

### Frequency Converter Accessory

**Boonton Radio Corp.**, Boonton, N. J., now provides complete FM and AM signal coverage from 0.1 to 216 mc, with a new accessory Univerter, Type 207-A, which increases the usefulness of the BRC Type 202-B FM signal generator.

The Type 207-A Univerter, a frequency converter accessory having unity gain, is designed for use with the Type 202-B FM signal generator to provide frequency

(Continued on page 91A)

The New STAVER
MINI-SHIELD
TRADE MARK REG. AND PAT. PEND.

The shield that fits all Miniature Tubes

A flexible shield that snugly fits all miniature tubes because it compensates for all variations in tube dimensions. Mini-Shields are made for both T5½ and T6½ bulb tubes. Send for catalog sheet.

THE Staver COMPANY INCORPORATED
91 PEARL ST. • BROOKLYN 1, N. Y.
ULSTER 5-6303

## News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

*(Continued from page 90A)*

coverage from 0.1 to 55 mc. Since the Type 202-B FM Signal Generator covers a frequency range from 54 to 216 mc, this combination of instruments permits continuous coverage from 0.1 to 216 mc.

The 207-A is provided with a frequency increment dial calibrated in increments of 5 kc from plus 300 kc through zero to minus 300 kc. This permits making selectivity measurements on narrow-band receivers.

The rf output voltage across a 53-ohm load connected at the unity gain jack is continuously variable from 0.1 microvolt to 0.1 volt by means of the 202-B signal generator attenuator. The gain is constant within ±1 db over the entire frequency range of the instrument.

### New Capacitors

**Pyramid Electric Co.**, 1445 Hudson Blvd., North Bergen, N. J., has developed several new capacitors. The "hearing aid" type 65 PTR paper tubular is designed for 65°C applications at 150-v dc continuous. It will withstand a 5-second test of 300-v dc, and meets standard RMA humidity test.

The "Glasseal" is a miniature hermetically sealed paper tubular, assembled in metal tubes with glass-metal terminals. They meet the demands of high vacuum,

*(Continued on page 92A)*

Measurements Corporation
MODEL 58

UHF RADIO NOISE & FIELD STRENGTH METER

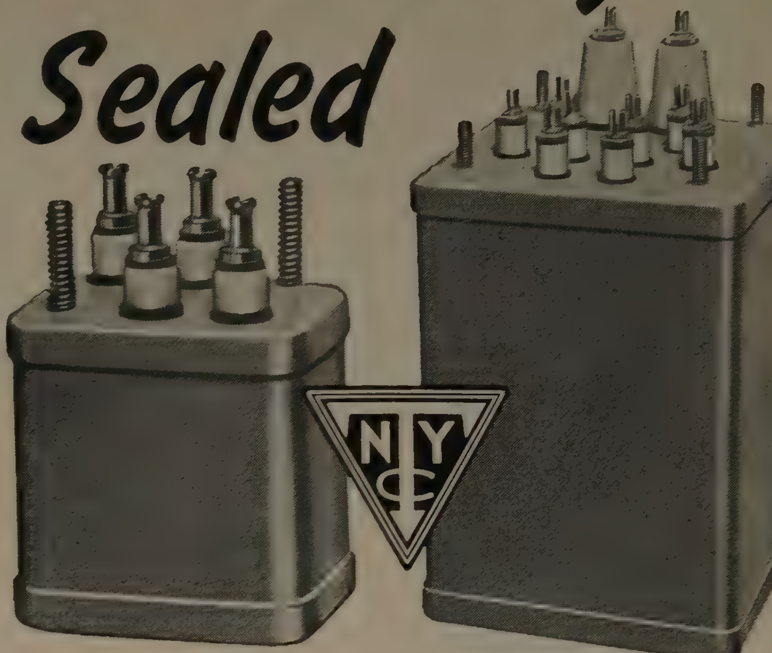
*Frequency Range*
15 Mc. to 150 Mc.

FREQUENCY ACCURACY: ±2%. Individually calibrated dial.

SENSITIVITY RANGE: 1 to 100,000 microvolts. Direct reading dial.

POWER SUPPLY: Built-in supply, 117 volts AC, 6 volts DC.

MEASUREMENTS CORPORATION
BOONTON NEW JERSEY



Hermetically Sealed

Power—Audio—Plate
TRANSFORMERS

FULL RANGE of STANDARD MIL-T-27 CASES

Prompt Delivery of Pre-production samples to your specifications

NEW YORK TRANSFORMER CO., INC.
ALPHA, NEW JERSEY

Important
SAVINGS
to VOLUME users
of small parts

..like these
SHOWN TWICE SIZE

thanks to
MULTI-SWAGE

If you need small tubular metal parts like these in large VOLUME, Bead Chain's MULTI-SWAGE Process can mean important savings to you.

**Much Cheaper Than Solid Pins**

Many prominent users of solid pins for electronic and mechanical purposes have cut costs by switching to Multi-Swaged *tubular* pins . . . without sacrificing strength or accuracy.

**Typical Applications—**

As terminals, contacts, bearing pins, stop pins, male-female connections, etc., in a wide variety of products such as Business Machines, Ventilator Louvres, Toys, Radio and Television Apparatus, Terminal-boards, Electric Shavers, Phonograph Pickups, etc.

Send part (up to ¼" dia. and to 1½" length) and your specs for a quotation or write for DATA BULLETIN.

THE BEAD CHAIN® MFG. CO.
11 Mountain Grove St., Bridgeport 5, Conn.

Manufacturers of BEAD CHAIN—the kinkless chain of a thousand uses, for pull and retaining chains and other industrial uses; plumbing, electrical, jewelry, fishing tackle and novelty products.

ALLISON
RADAR

NAVIGATIONAL AID
For MULTI-ENGINED

TRANSPORT AIRCRAFT
& HELICOPTERS

MILITARY—CIVIL

SMALL—COMPACT
LIGHTWEIGHT
EFFICIENT

LATEST MODELS:
E-2, ES-2, ESB-2
E-3, ES-3, ESB-3

- Sees Thunderstorm, Cores, Snow, Hail, Rain in advance.
- Promotes smoother flying for passenger safety and comfort.
- Warns of approaching aircraft in line of flight.
- Sees obstructions in true outline and/or direction.

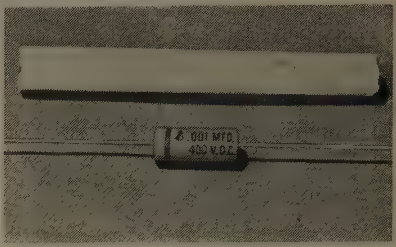
ALLISON
RADAR CORPORATION
11 W. 42 St., N.Y. 18 PEnn 6-5811-12

# News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

*(Continued from page 91A)*

high pressure, temperature cycling, immersion cycling and corrosion tests. Temperature range, −55° to +125°C. Capacitance range, 0.001 to 1.0 μfd. Voltage, 100 to 600 v dc operating.



Illustrated is the Type 85 LPT miniature paper tubular for 85°C application. This type is built in Phenolic-imprenated tubes, plastic end filled. Designed for continuous dc operation. The 85 LPT will withstand 2½ times rated voltage for 5 seconds.

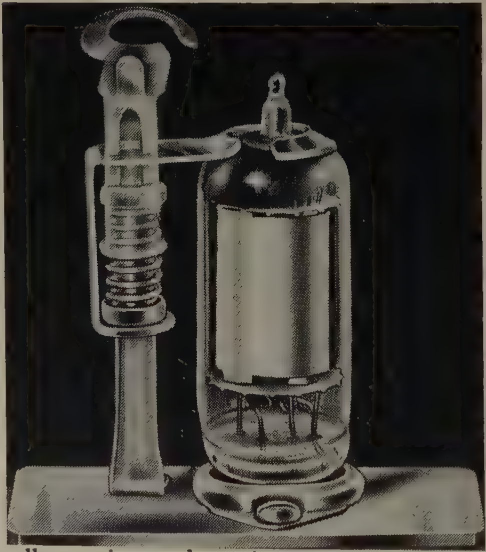
Also available is a complete 1951 catalog in color with many types and complete specifications listed.

## Recent Catalogs

• • • **Taylor Fibre Co.**, Norristown, Pa., has released a complete catalog listing all of the laminated plastics, rods, tubes, sheets, and fabricated parts which they have available at present to the industry.

*(Continued on page 94A)*

*New* BIRTCHER TUBE CLAMP
FOR MINIATURE TUBES



POSITIVE PROTECTION
AGAINST LATERAL AND
VERTICAL SHOCK!

The *New* Birtcher Type 2 Tube Clamp holds miniature tubes in their sockets under the most demanding conditions of vibration, impact and climate. Made of stainless steel and weighing less than ½ ounce, this *New* clamp for miniature tubes is easy to apply, sure in effect. The base is keyed to the chassis by a single machine screw or rivet...saving time in assembly and preventing rotation. There are no separate parts to drop or lose during assembly or during use. Birtcher Tube Clamp Type 2 is all one piece and requires no welding, brazing or soldering at any point.

If you use miniature tubes, protect them against lateral and vertical shock with the Birtcher Tube Clamp (Type 2). Write for sample and literature.

The BIRTCHER Corporation
5087 HUNTINGTON DRIVE • LOS ANGELES 32

IN BUSINESS LENZ SINCE 1904

# Wire for Defense Projects!

## HOOK-UP WIRES and CABLES for COMMUNICATIONS EQUIPMENT and ELECTRONIC INSTRUMENTS

FOR almost a half century Lenz has been producing insulated Wires and Cables for the Communications Industries, wires that are engineered and designed under high quality standards.

Now, Lenz is prepared to furnish hook-up wire and cables for defense projects, conforming to Government specification JAN-C-76 TYPES WL, SRIR and SRHV.

These Thermo-Plastic Insulated Wires, with or without Lacquered Braids, and the cables constructed of same, are available for use in Communications Equipment and Electronic Instruments.

For a dependable source for your wire and cable requirements, consult Lenz.

CONFORMING TO
**JAN-C-76**
TYPES WL, SRIR and SRHV

ON THE SEA

IN THE AIR

ON LAND

**LENZ ELECTRIC MANUFACTURING CO.**
1751 No. Western Ave. • Chicago 47, Illinois

LENZ

specify Sorensen RANGERS

(FULL-RANGE-VARIABLE DC SUPPLYS)

This versatile supply is a combination of the widely used Sorensen NOBATRON and a filter-variable output circuit. The result gives a continuously variable output voltage regulated against line and load changes through the full range of the instrument. Look at the specifications tabulated below—check them against your requirements. Where range of output, adaptability to diverse applications is essential, the Sorensen RANGER may well be your instrument of choice.

ELECTRICAL SPECIFICATIONS

| Model No. | SR-10 | SR-30 | SR-50 |
|---|---|---|---|
| Output range | 0 - 135 VDC | 0 - 30 VDC | 0 - 13 VDC |
| Current range | 1 - 10 amps | 3 - 30 amps | 5 - 50 amps |
| Input voltage | 95 - 130 VAC, 50 - 60 cycles, single phase | | |
| Regulation accuracy | ± 0.25 percent at any voltage setting from 3 VDC to top rating | | |
| Ripple | RMS max. 1% of output setting | | |

Meters — standard. Coarse and fine adjustment available.

Write for complete information

For other regulated DC supplys, investigate Sorensen's line of NOBATRONS (low-voltage) and B-NOBATRONS (high voltage).

POWER controlled converted Sorensen and company, inc.
375 FAIRFIELD AVE. • STAMFORD, CONN.

# News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 92A)*

## A Survey of Advertising Preferences of Members of The Institute of Radio Engineers

In December, 1950, the Advertising Department of the Institute initiated a survey to determine what advertisements were most acceptable and helpful to the membership—and why.

To obtain the information a Post Card (see illustration) was mailed to a portion of the membership representing a good cross-section of the country. This will be done each month, until the entire roster has been polled.

The reason why a member liked an advertisement fell into the following "pattern" (not necessarily in order of preference).

1. Specifications or Data.
2. Prices Quoted.
3. Application Data.
4. Engineering Drawing, Charts, Graphs, or Wiring Plan.
5. Good Visual Display—Attracts Attention.
6. Brief Yet Complete.
7. Dignified Presentation.
8. Generally Informative or Instructive.
9. Performance Characteristics.
10. No Ballyhoo or Advertising Blurbs.
11. No Answer.
12. Definitely Negative.

Other Reasons Infrequently Given—

- A. New Techniques.
- B. Order Blank in Ad.
- C. Physical Dimensions.
- D. New Product.
- E. Cut Away Drawing or Exploded View.
- F. Invites Correspondence.

The recipient of the card is requested to make a choice for the specific month, and then to state the reason for selection. Next, to choose an advertisement from the IRE Directory. Thirdly to state his type of business; and last to tell us if (A) he makes purchases, (B) sets specifications, (C) can influence buying, (D) has no interest in buying. Each of the last group is given a numerical value which is totaled to obtain a "point evaluation" of the men who prefer the ad.

*(Continued on page 97A)*

***This is a Continuing Survey from Month to Month***

***See Page 97A***

# News—New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 94A)*

This is the Result of the March Study, less than 3 votes not recorded.

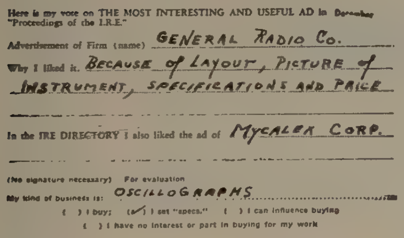
| Firm & Page | Vote | Ans. | Value |
|---|---|---|---|
| Hewlett-Packard Co. Pages 13A, 14A, 15A | 18 | 1 & 8 | 47 |
| General Radio Co. Cover IV | 15 | 1 | 44 |
| A. B. DuMont Labs., Inc. Pages 33A to 40A | 14 | 5 | 46 |
| United Transformer Co. Page 65A | 9 | 1 | 21 |
| Centralab Pages 81A, 82A, 83A | 6 | 8 | 17 |
| Fairchild Camera & Instr. Page 6A | 5 | 8 | 14 |
| General Electric Co. Page 3A | 4 | 6 | 7 |
| General Ceramics & Steatite Page 53A | 4 | 1 | 9 |
| Amperex Electr. Corp. Cover II | 4 | 8 | 7 |
| Cannon Elec. Dev. Co. Page 151A | 3 | 1, 5, 8 | 9 |
| Erie Resistor Corp. Page 97A | 3 | 1, 6, 10 | 7 |
| Intl. Resistance Co. Page 26A | 3 | 8 | 6 |
| Raytheon Mfg. Co. Page 71A | 3 | 5 | 7 |
| Simpson Electric Co. Page 41A | 3 | 1, 2, 5 | 7 |

Firms with single votes not recorded for Directory list.

| | | |
|---|---|---|
| General Radio Co. Cover IV | 11 | IRE Directory |
| Hewlett-Packard Co. Pages 236A & 237A | 6 | IRE Directory |
| Raytheon Mfg. Co. Catalog Section | 6 | IRE Directory |
| Mycalex Corp. of Amer. Catalog Section | 4 | IRE Directory |
| Centralab Catalog Section | 4 | IRE Directory |
| Approved Electr. Instr. Catalog Section | 4 | IRE Directory |
| RCA, Harrison Page 4A | 3 | IRE Directory |
| Ballantine Labs., Inc. Page 185A | 2 | IRE Directory |
| Collins Radio Co. Pages 226A & 227A | 2 | IRE Directory |
| Sylvania Elec. Prods. Page 62A | 2 | IRE Directory |

NOTE: *Vote* indicates total votes.
*Ans.* indicates the answer most frequently given, or in a tie, both.
*Value* indicates point evaluation total of voters.
*Directory* refers to the IRE Directory.

POST CARD



Here is my vote on THE MOST INTERESTING AND USEFUL AD in December "Proceedings of the I.R.E."

Advertisement of Firm (name) GENERAL RADIO CO.

Why I liked it. BECAUSE OF LAYOUT, PICTURE OF INSTRUMENT, SPECIFICATIONS AND PRICE

In the IRE DIRECTORY I also liked the ad of MYCALEX CORP.

(No signature necessary) For evaluation

My kind of business is: OSCILLOGRAPHS

( ) I buy; (✓) I set "specs." ( ) I can influence buying

( ) I have no interest or part in buying for my work

More than one reason was given by most men as the basis for selecting an ad. To avoid making a prejudiced selection, the researcher took the first reason in 90 per cent of the answers, on the basis that the most important factor would be stated first. This may be a fallacy, however, interpretation was avoided, in an attempt to be factual.

*(Continued on page 98A)*

Specify BREEZE "Monobloc" Waterproof and Pressure Sealed CONNECTORS

ANOTHER
BREEZE MARK
PRODUCT

The only APPROVED Monobloc System for Advanced Radar, Communications, and Electronic Equipment

Breeze "Monoblocs", with single piece plastic inserts, offer outstanding advantages in assembly, wiring, mounting and service in the field.

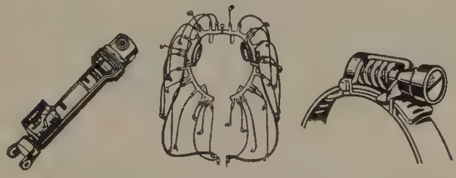
**Single piece inserts** make a tighter unit, eliminate the air spaces within conventional multiple-piece inserts, greatly reduce the opportunity for moisture shorts.

**Removable contact pins** make possible bench soldering of leads, quick, error free assembly of Breeze Waterproof Connectors and panel-type "Monobloc Miniatures."

**Single-Hole Panel Mounting** is all that is required for either Waterproof or Pressure Sealed types.

**Pressure Sealed** types are available for values up to and including 75 psi, or they can be specially engineered for greater pressures. They meet specified requirements of shock, vibration, salt spray, humidity and temperature cycling from −65° to +185° F.

Breeze "Monobloc" Waterproof and Pressure Sealed Connectors are engineered to your requirements in aluminum, brass or steel—in all sizes and capacities. They are fully tested and approved... cost no more than ordinary types.



*Other Breeze Precision Products*

**ACTUATORS:** All types, sizes. Complete control systems engineered to requirements. Above: Landing gear actuator Fairchild Packet.

**RADIO SHIELDING:** For any type of high or low tension system. New type "unit leads" or re-wirable leads. Flexible shielded conduit.

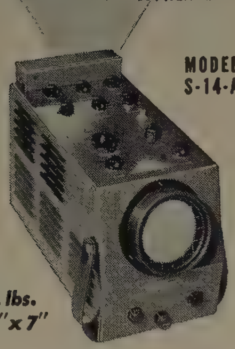
**"AERO-SEAL"** Worm-Drive Hose Clamps. Vibration proof, uniform clamping, use again and again. All clamps have stainless steel bands.

*Write for Details*

*If you have a tough connector problem, ask BREEZE for the answer!*

BREEZE
CORPORATIONS, INC.
41G SOUTH SIXTH ST., NEWARK, N.J.

THE
HI-GAIN Industrial
POCKETSCOPE
® BY WATERMAN



MODEL
S-14-A

Wt. 12¾ lbs.
12″ x 5¾″ x 7″

Another Waterman POCKETSCOPE providing the optimum in oscilloscope flexibility for analyses of low-level electrical impulses. Identified by its hi-sensitivity and incredible portability, S-14-A POCKETSCOPE now permits "on-the-spot" control, calibration and investigation of industrial electronic, medical and communications equipment. Direct coupling without peaking, used in the identical vertical and horizontal amplifiers, eliminates undesirable phase shifting. Designed for the engineer and constructed for rough handling, the HI-GAIN POCKETSCOPE serves as an invaluable precision tool for its owner.

Vertical and horizontal channels: 10mv rms/inch, with response within —2DB from DC to 200KC and pulse rise of 1.8μs. Non-frequency discriminating attenuators and gain controls with internal calibration of trace amplitude. Repetitive or trigger time base, with linearization from ½cps to 50KC with ± sync. or trigger. Trace expansion. Filter graph screen. Mu metal shield. And a host of other features.

WATERMAN PRODUCTS CO., INC.
PHILADELPHIA 25, PA.
CABLE ADDRESS: POKETSCOPE

WATERMAN PRODUCTS INCLUDE:
S-10-B GENERAL POCKETSCOPE
S-11-A INDUSTRIAL POCKETSCOPE
S-14-B WIDE-BAND POCKETSCOPE
S-15-A TWIN TUBE POCKETSCOPE
S-21-A LINEAR TIME BASE

Also RAKSCOPES, LINEAR AMPLIFIERS, RAYONIC® TUBES and other equipment

WATERMAN PRODUCTS

# News—New Products

(Continued from page 97A)

## Beta Gamma Survey Meter

The model RRCX-1 covers a range from 100 to 500,000 mr/hr using a null circuit with a visual indicator. This instrument can be manufactured at roughly a cost of $10 per unit in quantity. The complete unit is rugged and serviceable, making it suited for Civil Defense purposes. By the addition of a capacitor, the instrument can be used as a dosimeter. Experimental models are available to interested manufacturers from the designer **Radiation Research Corp.**, International Airport, West Palm Beach, Fla.

## Industrial Servo Amplifier

A servo amplifier has been placed in production for industrial use by **Servomechanisms, Inc.**, Post and Stewart Aves., Westbury, L. I., N. Y.

It is a small plugin unit, in which all the electronic elements required for one function in the control system are packaged together. Additional plugin facilities within the unit, comprising the transmission control networks, permit the adaptation of the amplifier to varying requirements for gain, phase shift, and damping.

Functional packaging of the amplifier elements simplifies maintenance, since a unit which gets out of order can be easily unplugged and replaced with another.

The SA-203 measures 5 inches in width ×8 inches in length ×4⅜ inches in height, and weighs 5½ pounds.

(Continued on page 101A)

WANTED
Your Surplus Laboratory Equipment, Tubes and Parts

Particularly items of good quality—
Boonton, Ferris, L & N, G-R, etc.

Also MILITARY SURPLUS items, such as ARC-1, ARC-3, ART-13, APR-1, APR-4, APR-5A, BC 348, etc. Transmitters, Receivers and components; TS-323, TS-173, TS-174, TS-175, BC-221, LAE, LAF, LAG Frequency Meters; TS-13, TS-14, TS-34, TS-35, TS-45, TS-146, TS-155, TS-263, BC-1277, all kinds "TS-," "IE-," and other Test Equipment needed, particularly for the Microwave Region; Spectrum Analyzers, etc.; also all kinds of quantity stocks, spare parts, plugs and cable.

We Buy for Cash, or Will Exchange for Needed Equipment

EDUCATIONAL INSTITUTIONS

and others who obtained donations of surplus equipment from the government: you are now legally free to sell, or exchange this equipment with us for items you can use. Sell NOW, during this period of temporary shortage, and aid the defense effort.

Please Describe and Price Your Items in Your First Letter

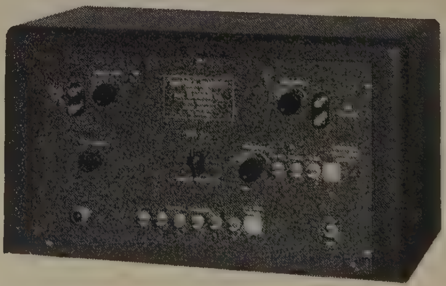
We specialize in reconditioning and assembling complete sets of equipment for use in the nation's laboratories and factories. We can, for instance, supply complete APR-4 receiving equipment, with continuous coverage of the frequency range 38-4,000 MC.; technical details and quotations furnished upon request.

Engineering Associates
434 Patterson Road
Dayton 9, Ohio WAlnut 1662

KAY

A Video Sweeping Oscillator With Frequency Marks

NEW NEW



THE MARKA-SWEEP MODEL VIDEO

The Marka-Sweep Model Video is an electronically swept video sweep oscillator covering frequencies up to 20 mc. Crystal positioned marks of pip type are provided at 1, 2, 5, 10, 15 and 20 mc. By use of external signal generator a variable frequency pip type marker is available. Output maximum is 0.3 volt from a 72 ohm internal impedance.

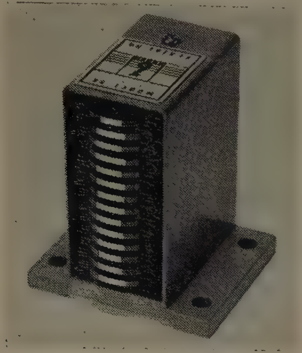
- Sweep Ranges: 50 kc to 10 mc and 50 kc to 20 mc.
- Crystal positioned pip type MARKS connected directly to oscilloscope.
- Separate attenuators on pip markers and video output.
- Sawtooth sweep for sweeping video output and deflecting oscilloscope horizontal.
- Produces Zero Level Baseline on oscilloscope.
- Price: $495.00 F.O.B. Factory.

KAY
KAY ELECTRIC COMPANY
14 Maple Avenue Pine Brook, New Jersey
Phone CAldwell 6-4000

(Continued from page 98A)

### Multichannel Magnetic Record-Reproduce Head

A new multichannel magnetic record-reproduce head has just been announced by **The Brush Development Co.,** 3405 Perkins Ave., Cleveland 14, Ohio. The BK-1500 series is available from 3 to 14 channels. Design features include balanced magnetic construction with gaps at front and rear. All gaps are in perfect alignment, and individual channels are cast into one integral block of specially selected synthetic resin. The complete assembly is nonmicrophonic and impervious to moisture. Shields and pole pieces are grounded to the housing. The head has a channel width of 0.050 inch with center to center spacing between channels of 0.125 inch. The gap is 0.0005 inch. Maximum output level at 1,000 cps and 7.5 inches per second ape speed is 0.002 volt rms.



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and production
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BASE, RARE AND PRECIOUS
METALS AND ALLOYS

SMALL UNITS
SMALL SIZES
CLOSE TOLERANCES

Nickel alloy, filament wire and ribbon: flat, grooved, crowned.

Grid wire electroplated.

Alloys for special requirements, bare or enamelled.

•

*Further details upon request.*

SECON
METALS CORPORATION
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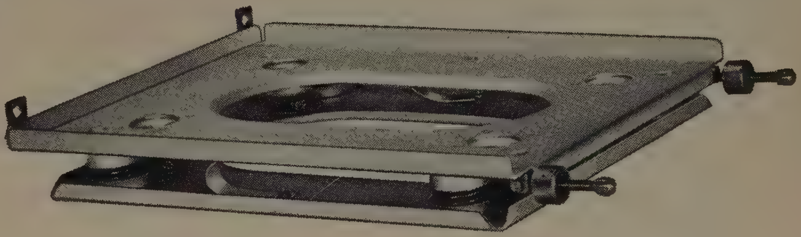
MUrray Hill 7-1594

ROBINSON ENGINEERED
MOUNTING
SYSTEMS

*Vibration isolation and shock protection for Airborne Equipment*



FOR LEAR INC....

Robinson Engineered Mounting Systems, similar to the Met-L-Flex base above, are being used by Lear Inc. for their famous aircraft electronic equipment after extensive in-flight and laboratory tests.

*First* Engineered Mounting Systems
• with *All-Metal*
• with MET-L-FLEX
• with APPROVAL

Vital and costly equipment aboard new high performance aircraft must be protected through extreme conditions of vibration, overload and shock. Engineers have found that "off the shelf" mounts seldom deliver the required results and therefore they look to Robinson Aviation for systems *engineered* for the specific application. Current Robinson Met-L-Flex Mounting Systems exceed the requirements of JAN-C-172A and Applicable specifications. Met-L-Flex systems are available in JAN form factors and special designs, to fit your equipment.

ROBINSON AVIATION INC.
TETERBORO, NEW JERSEY
*Vibration Control Engineers*

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*(Continued on page 103A)*

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M. D. Ercolino, Pres.
H. F. ANTENNA SPECIALISTS
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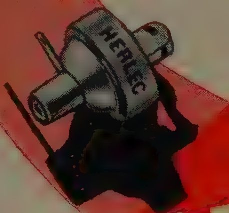
# Hermetically-sealed CUP CERAMIC CAPACITORS for Precision tuned Circuits



**With**
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**HERLEC**
**Dependability**

Top-notch capacitance stability, high Q, and excellent retrace characteristics are only part of the story on Sprague-Herlec Cup Ceramic Capacitors.

In addition, they are small, hermetically-sealed, and easy to mount securely against the effects of vibration and shock. Their extreme stability and compactness make them unexcelled for rigid frequency control applications and as reference capacitance standards in either laboratories or electronic circuits. Low self-inductance likewise makes them valuable in v-h-f bypass applications.

*With Sprague-Herlec metal cup ceramic capacitors in precision circuits, it is often possible to control the capacitance tolerance within ± 1% and the temperature coefficient tolerance within ± 10 parts per million!*

**For complete details write for Engineering Bulletin 603 to either Sprague or Herlec**

**HERLEC CORPORATION**

**422 N. 5th Street, Milwaukee, Wisconsin;** A wholly-owned subsidiary of the

**SPRAGUE ELECTRIC COMPANY**

**North Adams, Massachusetts**

Proceedings of the I.R.E. June, 1951, Vol. 39, No. 6. Published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers $1.00; non-members $2.25. Yearly subscription price: to members $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

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Chicago 1, Illinois

## *Important Meetings!*

●

**June 20-22**
**IRE 7th Regional Conference, Seattle, Washington**

Information: Chairman, Mr. A. E. Harrison, University of Washington Campus, Seattle, Washington.

●

**June 21-22**
**Electron Devices Conference, University of New Hampshire Durham, New Hampshire**

Information: Chairman, Prof. Herbert Reich, Dunham Laboratory, Yale University, New Haven, Conn.

●

June 25-29
1951 Summer General Meeting of AIEE, Royal York Hotel, Toronto, Canada.

●

June 27-28
1951 Annual IAS Summer Meeting, 7660 Beverly Blvd., Los Angeles, California.

●

June 28-30
Institute of Navigation National Meeting, Hotel New Yorker, New York, New York.

●

**August 22-24**
**1951 IRE West Coast Convention and Pacific Electronic Exhibit, Civic Auditorium, San Francisco, California.**

Heckert Parker, Exhibits Manager
600 Beach St., San Francisco 9

●

September 10-14
Instrument Society of America Convention and Instrument Show
**(Houston IRE Section Participating)**
Civic Auditorium
Houston, Texas

Information: Mr. Richard Rimbach, Exec. Secr., 921 Ridge Ave., N.S. Pittsburgh 12, Pennsylvania

●

**October 22-24**
**1951 National Electronics Conference, Edgewater Beach Hotel Chicago, Illinois**

Exhibits: R. M. Soria, American Phenolic Corp., 1830 South 54th Ave. Chicago 50, Illinois

●

**October 29-31**
**Radio Fall Meeting, King Edward Hotel Toronto, Ont., Canada**

●

November 1-3
Audio Fair, Hotel New Yorker
New York City, New York

Exhibits: Mr. Harry Reizes, 342 Madison Ave., New York 17

●

***IRE Regional Meetings and Shows Speed Electronic Progress***

Complete Coverage!

# 2 to 700,000,000 cps

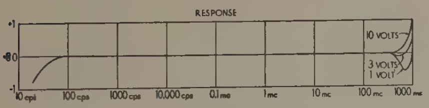
## -hp- precision voltmeters for every ac voltage measuring need!

From 2 cps to 700 mc, there's an accurate, easy-to-use *-hp-* voltmeter for any voltage measuring job. You can choose from 5 precision instruments (including a battery-operated portable unit) the dependable *-hp-* voltmeter that exactly fills your need. Each gives you familiar *-hp-* operating characteristics of high sensitivity, wide range, broad applicability, time-saving ease of operation. *-hp-* also provides a complete line of voltmeter accessories—voltage dividers, connectors, shunts and multipliers—to extend the useful range of your equipment. For complete details, see your *-hp-* sales representative or write direct.

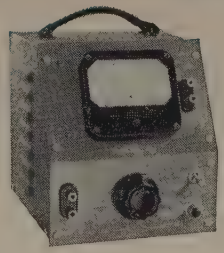
### New *-hp-* 410B Vacuum Tube Voltmeter

Gives same wide range and flat response performance as *-hp-* 410A voltmeter, but sets new standard of mechanical convenience, ease of operation, minimum bench space. Readily detachable probe leads fit in handy compartment in new, compact, streamlined case. Special diode probe design places capacity of approximately 1.3 $\mu\mu$fd across circuits under test. Shunt impedance is extremely high—10 megohms at low frequencies—thus circuits under test are not disturbed and true voltage readings are assured. New *-hp-* 410B provides 1 db accuracy from 20 cps to 700 mc; and may be used as a voltage indicator up to 3,000 mc. Also serves as audio or dc voltmeter or ohmmeter.

| INSTRUMENT | PRIMARY USES | FREQUENCY RANGE | VOLTAGE RANGE | INPUT IMPEDANCE | PRICE |
|---|---|---|---|---|---|
| *-hp-* 400A | General purpose ac measurement | 10 cps to 1 mc | .005 to 300v<br>9 ranges | 1 megohm<br>24 $\mu\mu$fd shunt | $185.00 |
| *-hp-* 400B | Low frequency ac measurements | 2 cps to 100 kc | .005 to 300v<br>9 ranges | 10 megohms<br>24 $\mu\mu$fd shunt | $195.00 |
| *-hp-* 400C | Wide range ac measurements<br>High sensitivity | 20 cps to 2 mc | .0001 to 300v<br>12 ranges | 10 megohms<br>15 $\mu\mu$fd shunt | $200.00 |
| *-hp-* 404A | Portable, battery operated | 2 cps to 50 kc | .0005 to 300v<br>11 ranges | 10 megohms<br>20 $\mu\mu$fd shunt | $185.00 |
| *-hp-* 410B | Audio, rf, VHF measurements; dc voltages; resistances | 20 cps to 700 mc | 0.1 to 300v<br>7 ranges | 10 megohms<br>1.3 $\mu\mu$fd shunt | $245.00 |



Response, *-hp-* 410B Voltmeter



### *-hp-* 400C Vacuum Tube Voltmeter

General purpose precision voltmeter offering wide range, high sensitivity, high stability. Quick-reading linear meter scale shows RMS volts or dbm direct from —72 dbm to +52 dbm. Broad usefulness includes direct noise or hum measurements, transmitter and receiver voltages, audio, carrier or supersonic voltages, or power gain. Also may be used as 54 db amplifier to increase signal level to oscilloscopes, recorders, power amplifiers, etc.

### *-hp-* 404A Battery-Operated Voltmeter

Precision vacuum tube instrument for general voltage measurement where ac power is not available. Compact, portable, splash-proof—ruggedly constructed for field operations. Wide voltage range permits all types of measurements including remote broadcast line and carrier checks, strain gauge system tests, telemetering and geophysical circuit measurements, etc. In the laboratory, offers completely hum-free measurements of very low noise level.

**HEWLETT-PACKARD CO.**

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Export: Frazar & Hansen, Ltd.

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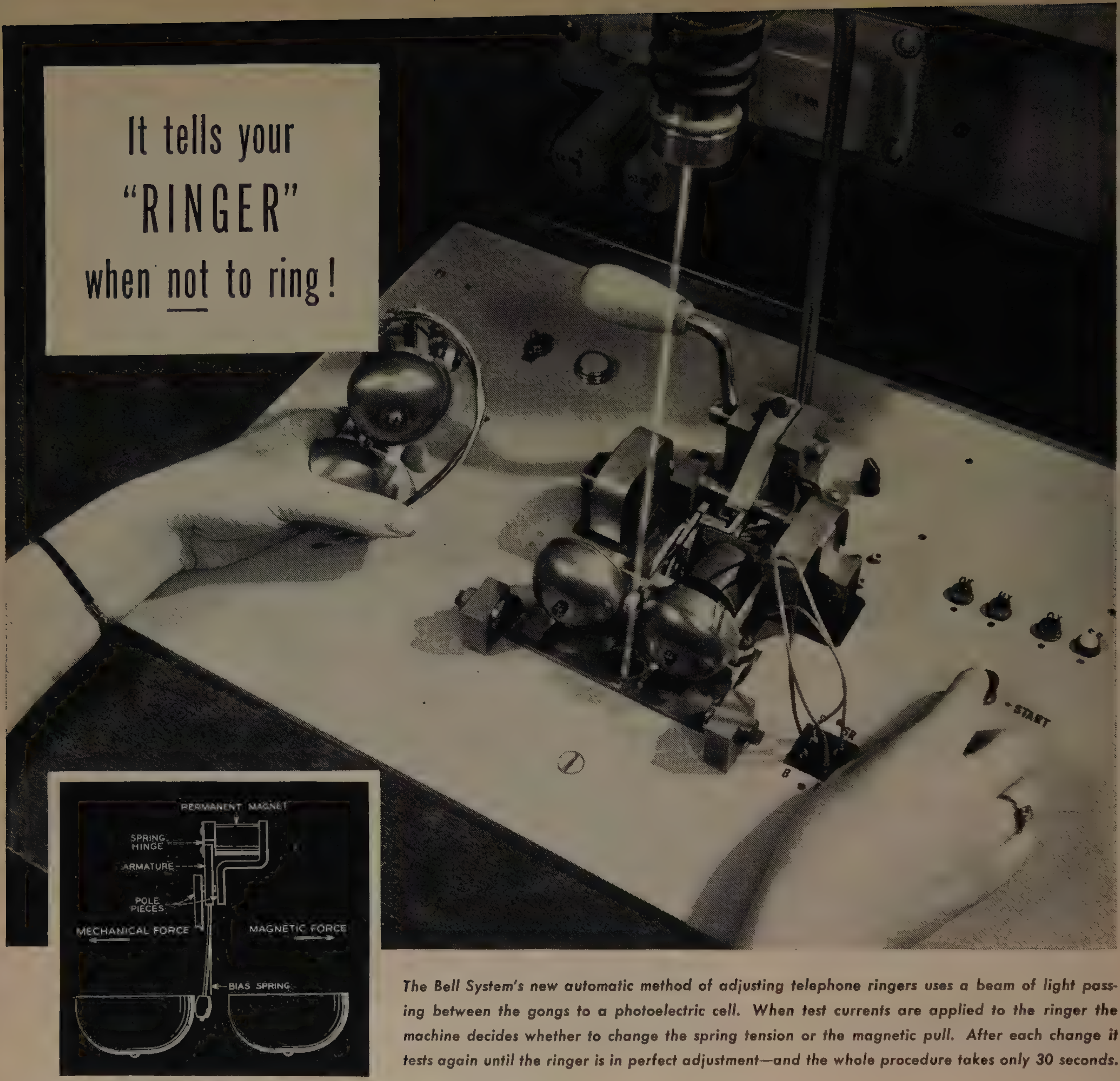
**HEWLETT-PACKARD -hp- INSTRUMENTS**

# It tells your "RINGER" when not to ring!



*The Bell System's new automatic method of adjusting telephone ringers uses a beam of light passing between the gongs to a photoelectric cell. When test currents are applied to the ringer the machine decides whether to change the spring tension or the magnetic pull. After each change it tests again until the ringer is in perfect adjustment—and the whole procedure takes only 30 seconds.*

To you, it's your familiar telephone bell. To telephone engineers, it's a "ringer." And it has two jobs to do. It must ring, of course, when someone calls you. And it must overlook the numerous electrical impulses which do not concern it, such as those sent out by your dial.

Ability to respond to some impulses, to ignore others, requires exact adjustment between the pull of a magnet and the tension of a spring. If they are out of balance your telephone might tinkle when it oughtn't, or keep silent when it should ring.

In the past, adjustment was made by hand, little by little until the proper setting was reached. It took time. But now Bell Laboratories engineers have developed a machine which adjusts new ringers perfectly, before they leave the Western Electric Company plants where they are made. And the operation takes just 30 seconds.

This is another example of how the Laboratories work constantly to improve every phase of telephony—keeping the costs low while the quality of service grows higher and higher.

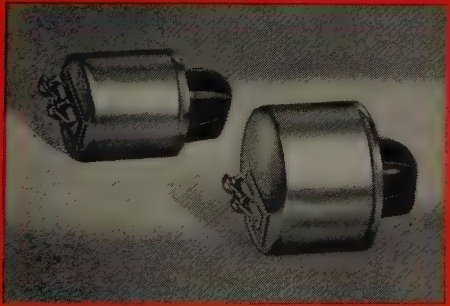
## BELL TELEPHONE LABORATORIES

*WORKING CONTINUALLY TO KEEP YOUR TELEPHONE SERVICE ONE OF TODAY'S GREATEST VALUES*

Ohmite Rheostats, in nine sizes from 25 to 750 watts, meet all requirements of Specification JAN-R-22.

# OHMITE RHEOSTATS

## MEET REQUIREMENTS OF JOINT ARMY-NAVY SPECIFICATION JAN-R-22



Models H (enclosed) and J (enclosed)
Also AN 3155 (AN-R-14a)

| TYPE | OHMITE MODEL | WATT RATING |
|---|---|---|
| RP10 | H | 25 |
| RP11 | H enclosed | 12.5 |
| RP15 | J | 50 |
| RP16 | J enclosed | 25 |
| RP20 | G | 75 |
| RP25 | K | 100 |
| RP30 | L | 150 |
| RP35 | P | 225 |
| RP40 | N | 300 |
| RP45 | R | 500 |
| RP50 | T | 750 |

### OHMITE RHEOSTATS MEET THESE RIGID TESTS:

- ★ 5-Hour Vibration Test (Required for RP 10-11-15-16-20-25)
- ★ 50-Hour Salt-Spray Corrosion Test
- ★ 150-Hour 95% Humidity Electrolysis Test

and other tests as prescribed in Specification JAN-R-22

By meeting these severe Joint Army-Navy requirements, Ohmite Rheostats have proved what industry has long accepted as true—that they can be depended upon for unfailing performance under the toughest operating conditions. All-ceramic construction . . . a smoothly gliding metal-graphite brush . . . uniform windings locked in place by vitreous enamel . . . insure close control throughout years of trouble-free service. It will pay you to standardize on Ohmite Rheostats for your product.

*Be Right with*

**OHMITE**®

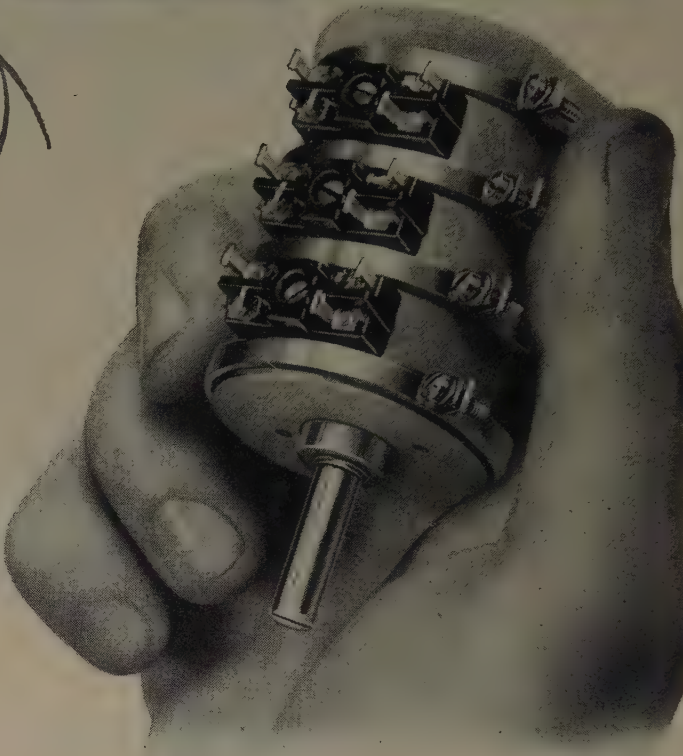
RHEOSTATS · RESISTORS · TAP SWITCHES

OHMITE MFG. CO., 4862 Flournoy St., Chicago 44, Ill.

# Here's Fairchild's Newest Potentiometer!

## TYPE-746
## PRECISION POTENTIOMETER OFFERS:

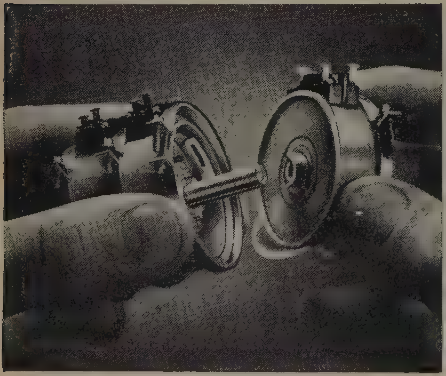
- Low Torque
- Accurate Phasing
- Quick Replacement
- Ganging up to 20 on a shaft



The finest we've ever built! That's our idea of the new "746". It's got lower torque, a new more accurate phasing adjustment, and a new method of ganging that makes it easy to put as many as twenty cups on a single shaft. Individual cups in a gang are easily replaced if necessary.

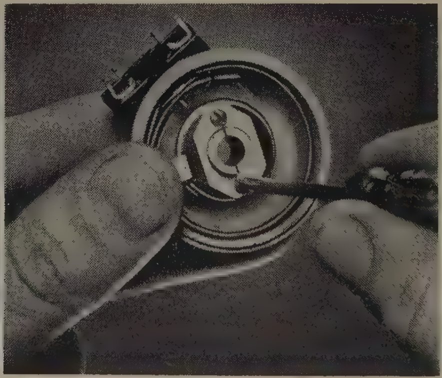
The new potentiometer is available with linear or non-linear windings to meet your specifications. Its attractive case is made of grey anodized aluminum.

The "746" is just one of the complete Fairchild family of precision potentiometers. What are your requirements? Write, giving details, to *Fairchild Camera and Instrument Corporation, 88-06 Van Wyck Boulevard, Jamaica 1, N.Y.* Dept. 140-13H1.



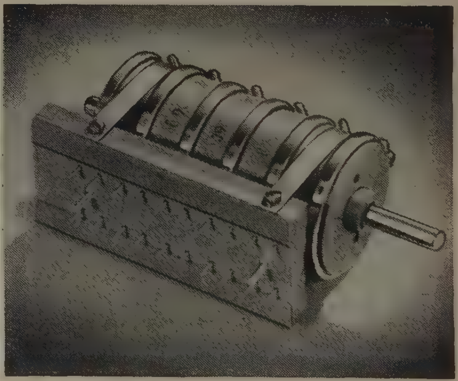
### EASY REPLACEMENT

To replace a unit in a "746" gang, loosen connecting-band screws, remove "cup," slip new "cup" under bands, and tighten screws. This feature pays off in experimental work where circuit elements are changed periodically.



### ACCURATE PHASING

A new type phasing adjustment is simpler and more accurate. A retainer plate clamps shaft to wiper arm. To adjust for phasing, loosen two screws, set the arm to the correct position, then tighten screws.



### FLEXIBLE DESIGN

Typical of the special consideration Fairchild gives to its customers' special requirements is this plug-in version of the "746." Where fast servicing is a must, the advantages of this "quick-change" unit are quite apparent.

### SPECIFICATIONS

**Accuracy (overall resistance)**—±.5% (linear), ±1.0% or better (non-linear)

**Mechanical accuracy**—
- concentricity (shaft to pilot)—.0015 in. FIR max.;
- radial play—.0009 in. FIR max.;
- shaft—centerless ground stainless steel to .2500 diam. (+.0000, −.0005 in.);
- pilot hub—machined to .5000 (+.0000, −.0005 in.)

**Torque**—1.5 oz-in.

**Dimensions**—diameter 1.750 max.; length (1 cup) .800 in. ±.009 in.; added length per unit ganged .580 in. ±.002 in.

**Case**—grey anodized aluminum

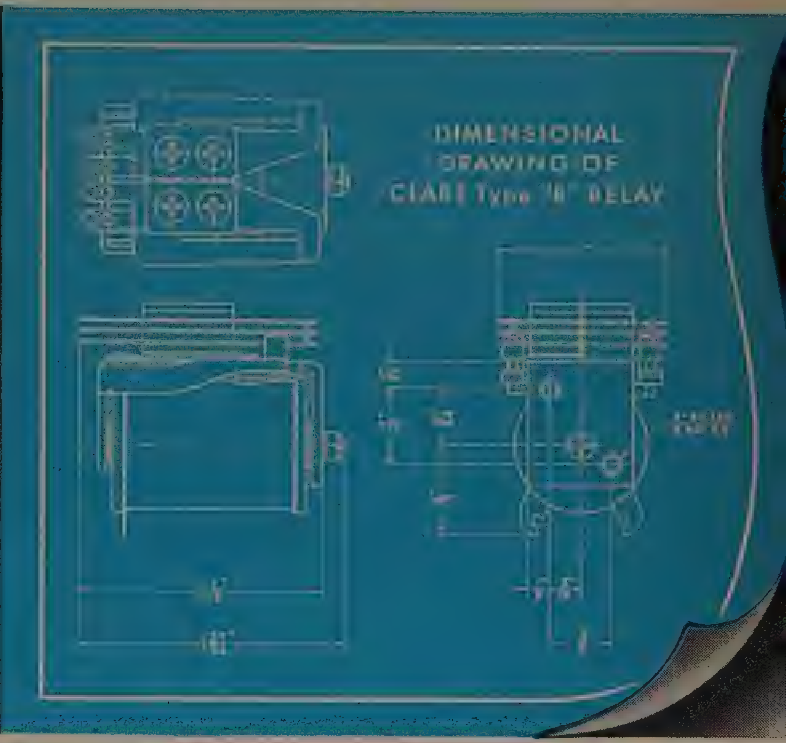
**FAIRCHILD**
**PRECISION POTENTIOMETERS**

# A NEW CLARE RELAY...

# the Type "R" combines extremely small size with unusual sensitivity and long life

CLARE Type "R" RELAY

Approximately Actual Size



## SPECIFICATIONS

**SIZE**
Length: 1 7/16"—Height: 1 3/64"—Width: 1"

**WEIGHT**
Approximately 2 ounces

**COIL**
Single or double-wound

**OPERATING VOLTAGE**
Up to 230 volts d-c

**ARMATURE**
Single or double arm

**CONTACT ASSEMBLY**
Form A to C. Maximum of 10 springs in each pileup.

**MOUNTING**
Two #4-40 tapped holes in end of heelpiece

● This new CLARE Type "R" d-c Relay embodies many features of the famous CLARE Type "K" Relay, which was the first to combine the advantages of a telephone-type relay with the small size, light weight and resistance to vibration required to meet the rigid demands of aircraft service.

In appearance, the Type "R" resembles the Type "K", but, through hardly noticeable structural differences, CLARE has given the new Type "R" even greater sensitivity and operating range. Both relays use the same contact springs, but the Type "R" coil is longer and of larger diameter, to provide greater winding space. Life expectancy of the new relay has been not only increased but multiplied.

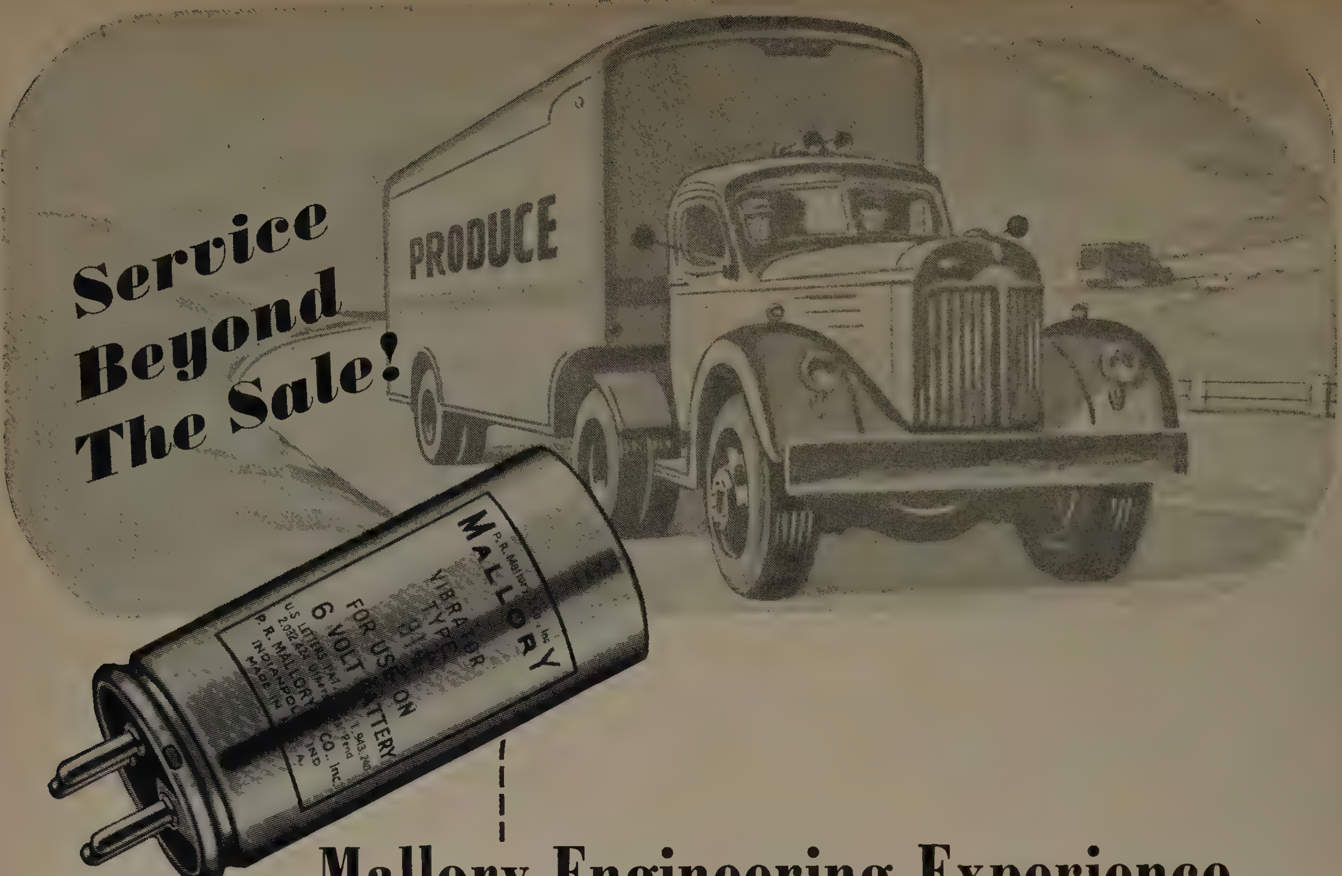
The CLARE Type "R" Relay retains in an improved form the reed armature suspension which discerning engineers have come to recognize as one of the subtler reasons for the superior performance of CLARE Type "K" Relays over other relays of comparable size and somewhat similar appearance.

The Type "R" is available as either an open or hermetically sealed relay. Clare sales engineers are located in principal cities to give you firsthand information on this new relay and to cooperate with you on any complex relay problem. Call them or write to C. P. Clare & Co., 4719 West Sunnyside Avenue, Chicago 30, Illinois. In Canada: Canadian Line Materials Ltd., Toronto 13. Cable Address: CLARELAY.

***Write for CLARE Bulletin No. 115***

**CLARE RELAYS**

*First in the Industrial Field*



# Mallory Engineering Experience

## *Solves Power Supply Problem With Unique Vibrator Application*

### MALLORY VIBRATORS

Mallory Vibrators are based on exclusive design and manufacturing methods that assure long, trouble-free service. Send the details of your application. Get Mallory's recommendation on the Vibrator or Vibrapack* power supply best suited to your needs.

**Reg. U.S. Pat. Off.*

The versatility of the Mallory Vibrator and the practicality of Mallory engineering have been demonstrated in a growing variety of power supply applications.

One customer had been experimenting with germicidal lamps in produce trucks to retard bacterial action...but was stumped by the need for an efficient power supply. Mallory tackled the problem and came up with an ingenious application of the Mallory Vibrator . . . plus complete technical data for producing the complete power unit assembly.

*That's service beyond the sale!*

Mallory electronic know-how is at your disposal. What Mallory has done for others can be done for you!

Vibrators and Vibrapack* Power Supplies

P.R. MALLORY & CO., Inc.
MALLORY

P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA

**SERVING INDUSTRY WITH**

Electromechanical Products
*Resistors* *Switches*
*TV Tuners* *Vibrators*

Electrochemical Products
*Capacitors* *Rectifiers*
*Mercury Dry Batteries*

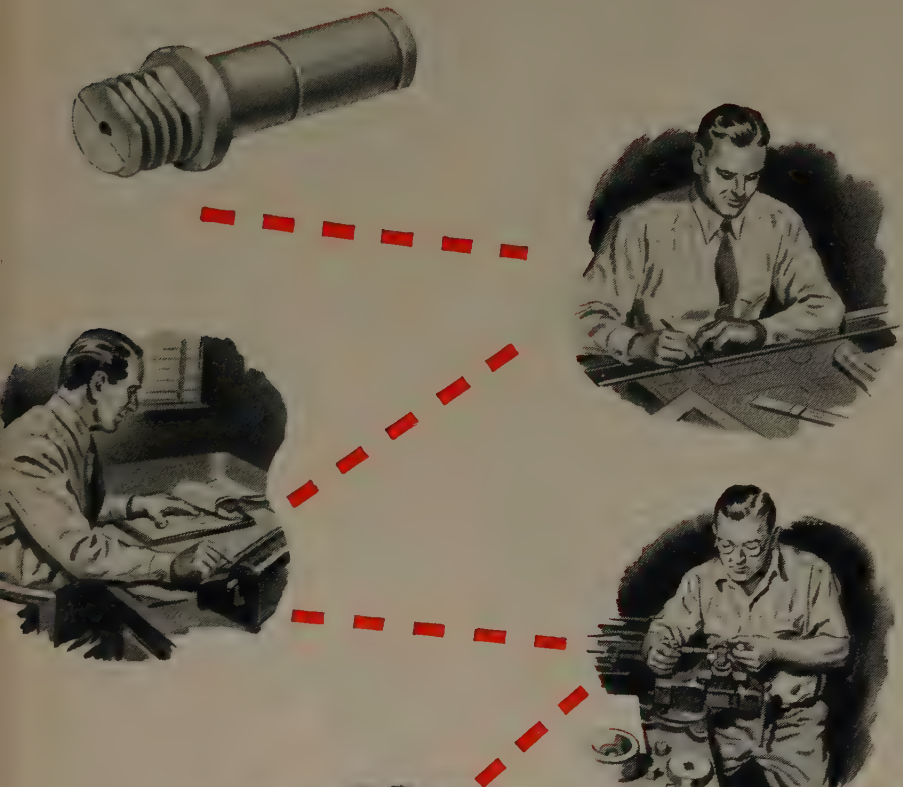
Metallurgical Products
*Contacts* *Special Metals*
*Welding Materials*

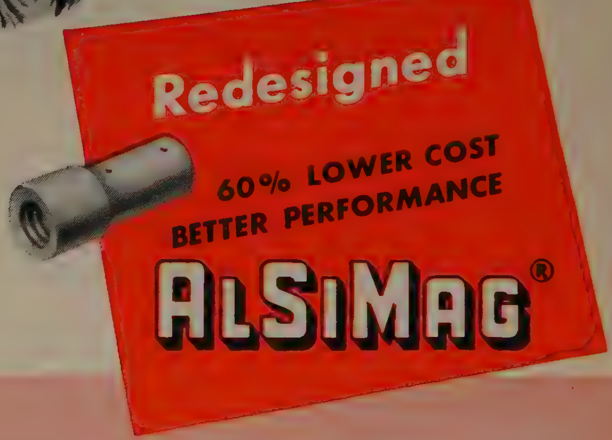
# It's a fact that

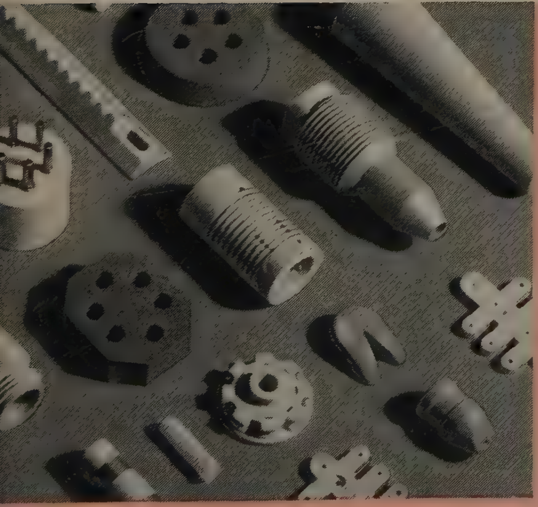
✓ Very few customers object to our suggestions on redesign. For those who object, we make their designs exactly as they want them. Most customers appreciate the close cooperation given and gladly use our fifty years of specialized experience.

*Here are the reasons:*

1. *Makes best use of favorable ceramic characteristics, avoids misapplication.*
2. *Designs are kept within practical manufacturing limits.*
3. *Subsequent assembly operations frequently are speeded.*
4. *Several components can sometimes be combined into one.*
5. *Maximum economy.*







Diversified equipment which is far beyond the reach of the average ceramic manufacturer permits us to use the best production method for the job.

Illustrated is a redesign which saved about 60% in original price, saved assembly time, and gave improved performance. Our engineers and production men will make recommendations on request. Send blue prints or samples with data on operating conditions and tolerance requirements. By working together we can serve you better.

**AMERICAN LAVA CORPORATION**

50TH YEAR OF CERAMIC LEADERSHIP

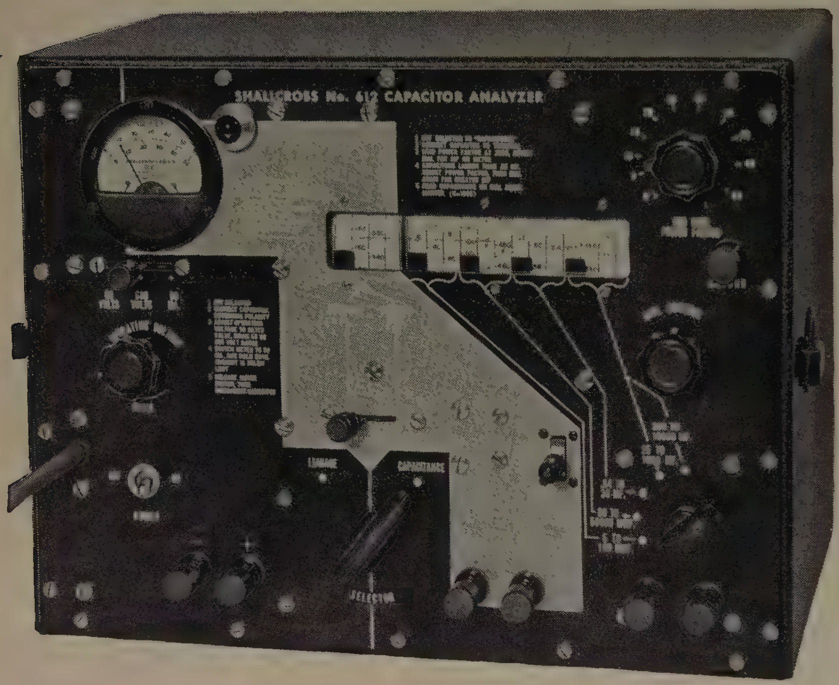
CHATTANOOGA 5, TENNESSEE

OFFICES: METROPOLITAN AREA: 671 Broad St., Newark, N. J., Mitchell 2-8159 • CHICAGO, 228 North LaSalle St., Central 6-1721
PHILADELPHIA, 1649 North Broad St., Stevenson 4-2823 • LOS ANGELES, 232 South Hill St., Mutual 9076
NEW ENGLAND, 38-B Brattle St., Cambridge, Mass., Kirkland 7-4498 • ST. LOUIS, 1123 Washington Ave., Garfield 4959

SHALLCROSS

612 CAPACITOR ANALYZER

ACCURATE • WIDE RANGE • DIRECT READING



Typical of the high quality, completeness and utility of Shallcross instruments, this analyzer provides a convenient means of determining the following capacitor characteristics:

| TEST | RANGE | ACCURACY |
|---|---|---|
| Leakage | 0-60 V d-c | ±3% of full scale |
| | 0-600 V. d-c | ±3% of full scale |
| | 0-50 MA. d-c | ±3% of full scale |
| Insulation | 1.1-120 megohms | ±20% |
| Resistance | 110-12,000 megohms | ±20% |
| Capacitance | 5-100 mmf | ±3 mmf |
| | 80-50,000 mmf | ±5% |
| | 0.04-30 mfd | ±5% |
| | 25-1,200 mfd | ±10% |
| | 250-12,000 mfd | ±20% |
| Power Factor | 0.50% | ±10% |

SHALLCROSS MANUFACTURING COMPANY

Collingdale, Pa.

PRECISION RESISTORS • D-C BRIDGES • DECADE RESISTANCES • HIGH-VOLTAGE MEASURING INSTRUMENTS • GALVANOMETERS • LOW-RESISTANCE TEST SETS • ROTARY SELECTOR SWITCHES • TRANSMISSION TEST SETS • CUSTOM-BUILT ELECTRONIC SPECIALTIES

# Something New

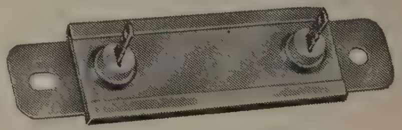
## NEW OVAL SELECTOR SWITCHES

Several new oval rotary selector switches are described in Bulletin L13 just issued by the Shallcross Manufacturing Co., Collingdale, Pa. Six basic plates and three rotor types produce switches having from one to three poles per deck or gang and with other desired mechanical and electrical details. As many as 18, 9 or 6 positions may be obtained in single-, double-, or triple-pole types respectively. These may be single-, double-, or triple-pole decks exclusively or a combination of different types.

## VERTICAL STYLE PRECISION RESISTORS FOR JAN USES

TYPE 140A
SHALLCROSS
MFG. CO.

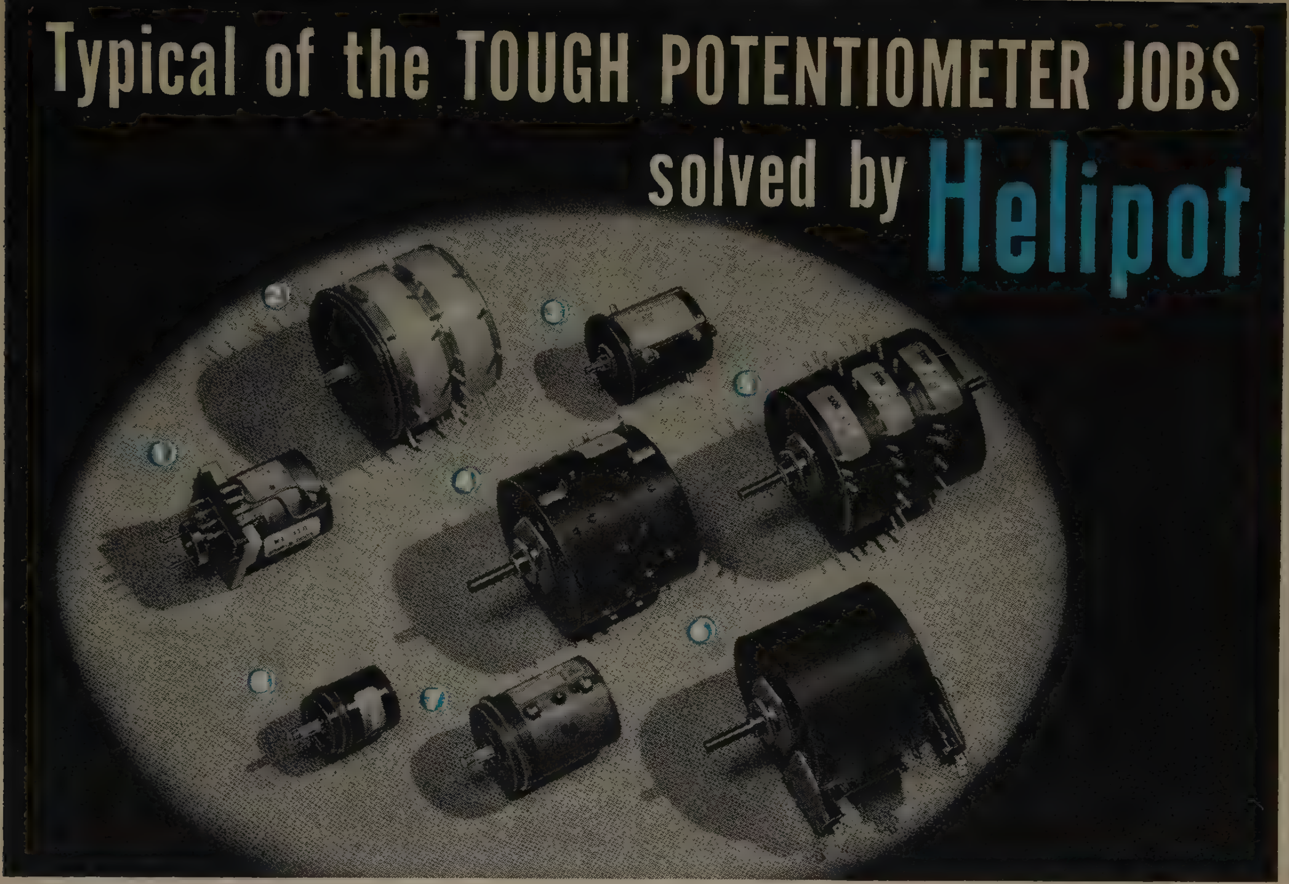
Improved vertical style precision wire-bound resistors for use where mounting requirements make it desirable to have both terminals at the same end of the resistor have been introduced by the Shallcross Manufacturing Co., Collingdale, Pa. These units provide a longer leakage path from the mounting screws to the terminals. Known as Shallcross Types BX120, BX140, and BX160, they are designed to meet JAN requirements for styles RB40B, RB41B and RB42B respectively. For commercial uses, the resistors carry somewhat higher ratings than for JAN applications. Wire leads instead of terminals can be furnished if desired. Complete details will gladly be sent on request to the manufacturer.



## FLAT, METAL-ENCASED WIRE-WOUND RESISTORS

Flat, metal-encased, Type 265A wire-wound power resistors introduced by the Shallcross Manufacturing Company, Collingdale, Pa., are space wound, have mica insulation, and are encased in aluminum for mounting flat against a metal chassis. At 175° C. continuous use they are conservatively rated for 7½ watts in still air and 15 watts when mounted on a metal chassis. Write for Bulletin 122.

# Typical of the TOUGH POTENTIOMETER JOBS solved by Helipot



## Precise Accuracy + Maximum Versatility + Space-saving Compactness

**The potentiometers illustrated above are typical examples of the tough problems HELIPOT engineers are solving every day for modern electronic applications. If you have a problem calling for utmost precision in the design, construction and operation of potentiometer units—coupled with minimum space requirements and maximum adaptability to installation and operating limitations—bring your problems to HELIPOT. Here you will find advanced "know-how," coupled with manufacturing facilities unequaled in the industry!**

**The HELIPOTS above—now in production for various military and industrial applications—include the following unique features . . .**

① This 10-turn HELIPOT combines highest electrical accuracies with extremes in mechanical precision. It features zero electrical and mechanical blacklash...a precision-supported shaft running on ball bearings at each end of the housing for low torque and long life . . . materials selected for greatest possible stability under aging and temperature extremes . . . special mounting and coupling for "plug-in" convenience . . . mechanical and electrical rotation held to a tolerance of ½° . . . resistance and linearity accuracies, ±1% and ±0.025%, or better, respectively.

② *This four-gang assembly of Model F single-turn potentiometers has a special machined aluminum front end for servo-type panel mounting, with shaft supported by precision ball bearings and having a splined and threaded front extension. Each of the four resistance elements contains 10 equi-spaced tap connections with terminals, and all parts are machined for greatest possible stability and accuracy.*

③ This standard Model A, 10-turn HELIPOT has been modified to incorporate ball bearings on the shaft and a special flange (or ring-type) mounting surface in place of the customary threaded bushing. This HELIPOT also contains additional taps and terminals at the ¼- and 9¾-turn positions.

④ *This standard Model B, 15-turn HELIPOT has a total of 40 special tap connections which are located in accordance with a schedule of positions required by the user to permit external resistance padding which changes the normally-linear resistance vs. rotation curve to one having predetermined non-linear characteristics. All taps are permanently spot-welded and short out only one or two turns on the resistance element—a unique HELIPOT feature!*

⑤ This six-gang assembly of standard Model F single-turn potentiometers has the customary threaded bushing mountings, and has shaft extensions at each end. The two center potentiometers each have 19 equi-spaced, spot-welded tap connections brought out to terminals. *Each tap shorts only two turns of .009" diameter wire on the resistance element.*

⑥ *This Model B, 15-turn HELIPOT has been modified to incorporate, at the extreme ends of mechanical and electrical rotation, switches which control circuits entirely separate from the HELIPOT coil or its slider contact.*

⑦ This 10-turn HELIPOT has many design features similar to those described for unit No. 1, plus the following additional features . . . a servo-type front end mounting . . . splined and threaded shaft extension . . . and a center tap on the coil. All components are machined to the highest accuracy, with concentricities and alignments held in some places to a few *ten-thousandths* of an inch to conform to the precision of the mechanical systems in which this HELIPOT is used. *Linearity accuracies frequently run as high as ±0.010%!*

⑧ *This single-turn Model G Potentiometer has been modified to incorporate a ball bearing shaft and a servo-type front end mounting. Special attention is given to contact designs and pressures to insure that starting torque does not exceed 0.2 inch-ounces under all conditions of temperature.*

**The above precision potentiometers are only typical of the hundreds of specialized designs which have been developed and produced by HELIPOT to meet rigid customer specifications. *For the utmost in accuracy, dependability and adaptability, bring your potentiometer problems to HELIPOT!***

**Representatives in all major areas of the United States. Export agents: Frathom Co., 55 W. 42nd St., New York 18.**

**THE Helipot CORPORATION, SOUTH PASADENA 6, CALIFORNIA**

# BUSINESS IN MOTION

## *To our Colleagues in American Business . . .*

Never has industry turned out more goods than during the past few years of unprecedented customer demands. In endeavoring to meet these, Revere has developed new techniques, established new plants, installed the newest equipment and modernized the old, and stepped up its training program, including the development of some new ideas in relation to safety. Throughout the country, similar steps have been taken by manufacturers generally. This is the response of free enterprise to the stimulus of a free and growing market. It is fortunate that American industry was not only willing but able to do this, because now it is evident that these facilities and these skills must be devoted more and more to the defense of our freedom.

NOTICE
MANUFACTURERS
WITH ORDERS RATED
D.O.

Defense Orders or "DO's" are being issued, and their volume is bound to increase. Already prime contractors are seeking sub-contractors, and sub-sub-contractors are receiving orders too, down to small local firms operating only a few machines. Perhaps few people realize the importance of the "small shop"; the fact is that these establishments have a tremendous total capacity supplementing that of the great corporations, which practically never make everything that is needed for a finished product such as a tank, a plane, a ship, radar equipment. The "smalls" are just as vital as the "bigs."

Revere knows that when the time of trial comes, it is more important than ever to increase production efficiency. This makes complete information essential to those who have taken on DO contracts. Revere pledges its full cooperation, and will gladly provide all it knows about its metals.

This knowledge is made available in two principal ways. First, there are many booklets containing technical data, including physical properties, and also in many cases suggestions as to recommended fabrication practices. In addition to the booklets, which are distributed on request, Revere either reproduces or summarizes them in the various Sweet's Files, Chemical Engineering Catalogue, Marine Catalogue, Refinery Catalogue. This printed material is therefore available freely to all who will ask for it, or look it up. The second way in which Revere's knowledge and skill is made available is through the Technical Advisory Service, a group of capable men whose collective experience covers practically all applications of copper and copper alloys, and aluminum alloys. In war and peace, these men have rendered invaluable service, collaborating closely on such matters as selection of the proper metal, temper, width, gauge, and in helping to solve production problems. As a result, scrap has been reduced, rejects lessened, production increased, money and materials saved. The services of the Technical Advisors are obtainable through the Revere Sales Staff which also has wide experience in the selection and application of Revere Metals. If you have orders whose specifications include non-ferrous metals, Revere will gladly place its information at your disposal.

If you purchase and work with other materials, Revere suggests that it should be realized that not only is American productive capacity tremendously greater, but that there has been a likewise large growth in knowledge about materials of all kinds. So it is recommended that no matter what you make now, or are called on to make in the future, you ask your suppliers to share their knowledge with you. It will make you and our country stronger.

**REVERE COPPER AND BRASS INCORPORATED**
*Founded by Paul Revere in 1801*
**Executive Offices:**
**230 Park Avenue, New York 17, N. Y.**
150th YEAR OF SERVICE TO AMERICA
**SEE "MEET THE PRESS" ON NBC TELEVISION EVERY SUNDAY**

EXTRA-DEPENDABLE

LONG-LIVED

WON'T SHORT

WITHSTAND HIGH HEATER-CYCLING

SHOCK-RESISTANT

# G-E 5-STAR ARINC TUBES ARE 5 WAYS BETTER!

**Take your design cue from airmen who praise the superiority of ARINC types!**

"Records kept by our company show replacement of ARINC tubes to be only 2% as compared with an average of 49% replacement for 13 types of regular tubes over the same period."

W. W. LYNCH, *System Communications Superintendent,*
Pan American World Airways System.

"Our company is using ARINC reliable tubes wherever possible. Experience has shown that equipment using these tubes seldom causes delays from tube failure."

J. H. CARMICHAEL, *President,*
Capital Airlines.

"The effect of using ARINC tubes in our equipment became immediately apparent. Off-schedule removals of airborne equipment due to tube failures have been materially reduced."

J. R. CUNNINGHAM, *Director of Communications,*
United Air Lines.

"During a six months' carefully controlled comparison service test, ARINC tubes required only ¼ as many replacements as first quality standard brand tubes."

FRANK R. WAGNER, *Supervisor of Radio, Electrical and Instrument Engineering,* TWA

"We have four VHF Navigation Receivers which have been in service for more than 1,000 hours each. Of the 104 ARINC tubes used, we have not had a single failure."

J. LANE WARE, *Supervisor of Communications Engineering,* National Airlines.

GL-5654

GL-5670

GL-5686

★★★★★ **A pioneer in manufacturing** ARINC tubes, G.E. offers nine 5-star types which set new standards of dependability and long life. Specify these tubes in electronic circuits now on your boards, to increase safety factor . . . reduce upkeep costs . . . *build reputation* for your equipment!

**G-E 5-star** ARINC tubes are a joint achievement of Aeronautical Radio, Inc., and General Electric Co. Built with exacting care—individually tested—they accent the reliability of altimeters, radio compasses, h-f receivers, other apparatus that guides air passengers safely to their destinations.

**Learn more about** G-E 5-star tubes! Write for Bulletin ET-B29A, which tells the story of their design, manufacture, and testing . . . describes and rates the various 5-star types . . . shows how *you*, as equipment builder or designer, can profit from their superior performance! *General Electric Company, Electronics Department, Schenectady 5, New York.*

GL-5654 —Sharp-cutoff r-f pentode
GL-5670 —H-f twin triode
GL-5686 —Power-amplifier pentode
GL-5725 —Semi-remote-cutoff r-f pentode
GL-5726 —Twin diode
GL-5749 —Remote-cutoff r-f pentode
GL-5750 —Pentagrid converter
GL-5751 —High-mu twin triode
GL-5814 —Medium-mu twin triode

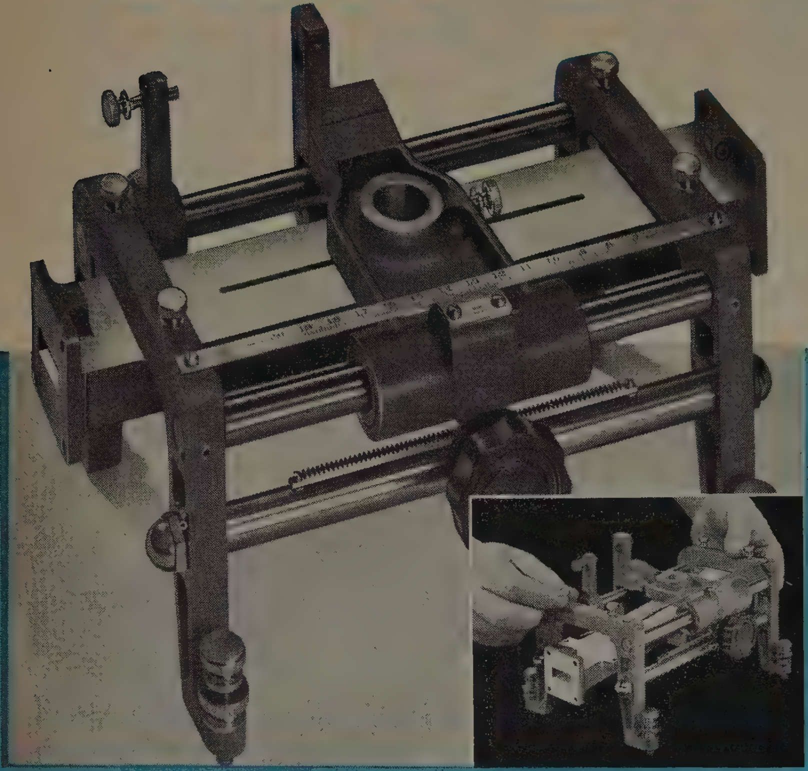
ENGINEERING AND APPLICATION DATA
Electronic Tubes
GENERAL ELECTRIC

GENERAL ELECTRIC

185-K4

# IMPEDANCE

## COMPLETE COVERAGE!



**–*hp*– 809B UNIVERSAL PROBE CARRIAGE with –*hp*– 810A WAVEGUIDE SLOTTED SECTIONS**

Now—a single probe carriage operates with up to 5 different slotted sections — waveguide or coaxial! This means important savings in time; lower investment in instrumentation. The new –*hp*– 809B Universal Probe Carriage mounts slotted sections covering frequencies from 3,000 to 12,400 mc (see table on opposite page)—and you can *interchange sections in 30 seconds* or less!

–*hp*– 809B Carriage is accurately calibrated in mm. for readings as low as 0.1 mm. Dial gauge may be readily mounted if more accurate readings are needed. Carriage travels on a new 3-point ball-bearing suspension system, and operates in conjunction with –*hp*– 442A Broad-Band Probe and –*hp*– 440A Coaxial Detector combination; or with –*hp*– 444A Untuned Probe. The extremely broad usefulness of this new Universal Carriage means far greater flexibility and lower cost for complete microwave instrumentation.

**CONTINUOUS microwave coverage, 10 mc to 12,400 mc. High mechanical stability. Simple operation. Broad applicability. Precision accuracy. Compact size!**

**New –*hp*– microwave equipment gives you *complete coverage* for VHF, UHF and SHF impedance measurements. Instrumentation includes VHF Bridges as well as the slotted coaxial and waveguide sections which are fundamental tools in impedance measurements. These instruments can be used to measure load or antenna impedance, system flatness, connector reflection, percentage of reflected power, standing wave magnitude or phase, characteristics of coaxial transmission lines or rf waveguide systems, characteristics of rf chokes, resistors, condensers.**

**For complete details see your –*hp*– sales representative or write direct.**

**HEWLETT-PACKARD COMPANY**

**2160D Page Mill Road • Palo Alto, California**

Sales representatives in principal areas.
Export: Frazar & Hansen, Ltd.
San Francisco, New York, Los Angeles

### –*hp*– 417A VHF DETECTOR

For use with –*hp*– 803A VHF Bridge. A super-regenerative (AM) receiver covering all frequencies 10 to 500 mc in 5 bands. Offers approx. 5 $\mu$v sensitivity over entire band; quick, easy operation, direct-reading frequency control. Thoroughly shielded, suitable for general laboratory use including approximate frequency checks, measurements of noise, interference, etc. $200.00 f.o.b. factory.

### –*hp*– 415A STANDING WAVE INDICATOR

Designed for use with all waveguide or coaxial slotted sections, to give direct reading of standing wave ratio in VSWR or db. Consists of high gain amplifier with low noise level, operating at fixed frequencies between 300 and 2,000 cps. (Normal frequency 1,000 cps., plug-ins for other frequencies available). Input circuits for use with crystal detector or bolometer. $200.00 f.o.b. factory.

**HEWLETT-PACKARD *hp* INSTRUMENTS**

# READINGS

# 10 to 12,400 mc.

### -hp- 805A/B COAXIAL SLOTTED SECTIONS

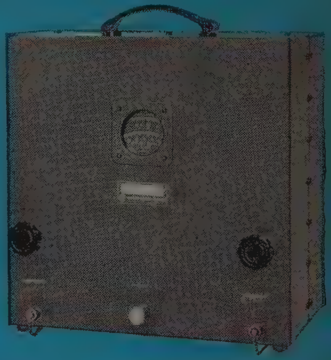
Continuous coverage 500 to 4,000 mc. High accuracy and mechanical stability; negligible slope, minimum leakage. Incorporates radically different structural design employing rigid parallel planes and a non-bowing central conductor. Probe setting readable in mm. to 0.1 mm. Maximum VSWR of basic section and connectors less than 1.04. *-hp-* 805A, 50 ohms impedance, for Type N connector and flexible cables. Model 805B, 46.3 ohms impedance, for ⅞″ rigid transmission lines.

### -hp- 806B COAXIAL SLOTTED SECTION

Continuous coverage 3,000 to 12,000 mc. Employs same time-tested parallel plane principle as *-hp-* 805A/B. Designed for use with *-hp-* 809B Universal Probe Carriage. Maximum VSWR of slotted section and connecters is 1.06 to 10,000 mc. Negligible slope, 50 ohm impedance. Uses Type N connectors for flexible coaxial cable. Sets new standard for mechanical stability in coaxial slotted sections.

### -hp- 803A VHF BRIDGE

Gives direct readings in impedance magnitude and phase, 10 to 500 mc. Rapid operation for new speed, convenience in reading impedance, or resistance and reactance. Operates on new principle of sampling magnetic and electric field of transmission line. Useful for comparative measurements, 5 to 1,000 mc. Impedance range 2 to 2,000 ohms. Phase angle —90° to +90°, at 52 mc and above. Offers utmost convenience in determining characteristics of antennas, transmission lines, rf chokes, resistors and condensers; in measuring connector impedances, standing wave ratios, percentage of reflected power, VHF system flatness.



### -hp- 440A COAXIAL DETECTOR

Tunable crystal and bolometer mount. May be used as an rf detector for coaxial systems between 2,400 and 12,400 mc. Fits Type N connectors; operates with bolometer or silicon crystal. $85.00 f.o.b. factory.

### -hp- 442A BROAD-BAND PROBE

May be used in combination with *-hp-* 440A to provide highly sensitive, easily tuned detector for slotted sections. Micrometer depth adjustment provides quick control of rf coupling. $75.00 f.o.b. factory.

### -hp- 444A UNTUNED PROBE

Frequency range 2,400 to 12,400 mc. Includes 1N26 silicon crystal. Highly sensitive, compact, easy to use. Requires no tuning. $50.00 f.o.b. factory.

### -hp- IMPEDANCE MEASURING EQUIPMENT

| INSTRUMENT | FREQUENCIES—COAXIAL | FREQUENCIES—WAVEGUIDE | PRICE (F.O.B. FACTORY) |
|---|---|---|---|
| 803A VHF BRIDGE | 10 to 500 mc | | $495.00 |
| 805A/B SLOTTED SECTION | 500 to 4,000 mc | | $475.00 |
| 806B SLOTTED SECTION* | 3,000 to 12,000 mc | | $200.00 |
| S810A SLOTTED SECTION† | | 2,600 to 3,950 mc | $450.00 |
| G810B SLOTTED SECTION* | | 3,950 to 5,850 mc | $ 90.00 |
| J810B SLOTTED SECTION* | | 5,850 to 8,200 mc | $ 90.00 |
| H810B SLOTTED SECTION* | | 7,050 to 10,000 mc | $ 90.00 |
| X810B SLOTTED SECTION* | | 8,200 to 12,400 mc | $ 90.00 |
| 809B UNIVERSAL PROBE CARRIAGE | For slotted sections, 3,000 to 12,400 mc | | $160.00 |

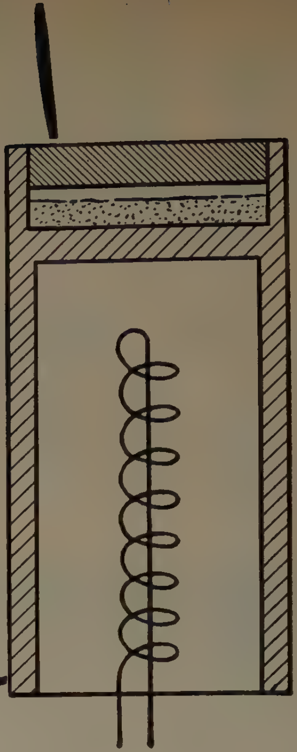
* Mounts in *-hp-* 809 B Universal Probe Carriage.

† Complete assembly including slotted section and carriage.

*Data Subject to Change Without Notice*

**HEWLETT-PACKARD INSTRUMENTS**

# *Announcing*
## the First Basically New Cathode in Over 25 Years
Providing High Emission Density with Long Life



# PHILIPS L-CATHODE

Illustrated above is the first basically new cathode in over 25 years—the new PHILIPS L-CATHODE. It offers these practical advantages to users of klystrons, disc seal triodes, iconoscopes, magnetrons, special cathode-ray tubes and other types where a high degree of reliability is essential:

1. *High emission current density with long life*
2. *Can be made in great variety of shapes and to close tolerances*
3. *Not damaged by high electrostatic forces*
4. *Reactivates after arcing*
5. *Can be reactivated after exposure to air*
6. *Withstands severe ionic and electronic bombardment*
7. *Maintains uniform emission characteristics during its life*

The PHILIPS L-CATHODE originated in the Philips Research Laboratories in Holland, and was further developed by Philips Laboratories, Inc., here. It was first described in the June 1950 *Philips Technical Review.* Leading electronic research laboratories have already used development types of various shapes with outstanding results.

Samples of PHILIPS L-CATHODES for experimental use can be supplied in limited quantities to those interested in producing greatly improved tubes of high emission density with longer life.

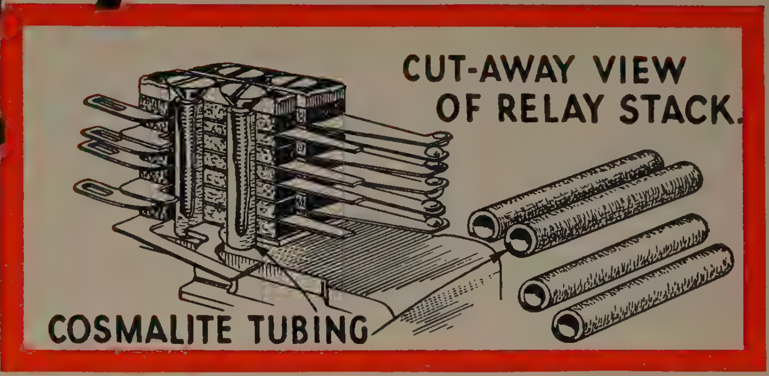
**PHILIPS LABORATORIES, INC.**

Dept. P6, Irvington-on-Hudson, New York

# ALLIED CONTROL RELAYS
*built with*
# *Cleveland* PHENOLIC TUBES

**ensure**

- ★ DEPENDABILITY
- ★ PRECISION PERFORMANCE
- ★ LONG LIFE
- ★ ECONOMY



The Allied Control Co. has built a long and enviable record as a quality supplier of control relays to both private industry and governmental services.

Their S K Relay shown above, is typical of the various Allied Relays in which CLEVELAND CONTAINER tubing provides excellent service.

It is likewise the answer for hundreds of other problems of manufacturers in the electrical industry.

## CLEVELITE* and COSMALITE*
### *Laminated Phenolic Tubing*

combines electrical and physical properties to meet the most exacting requirements.

CLEVELITE is produced in six grades . . .
COSMALITE in four grades . . .
A grade for EVERY need.

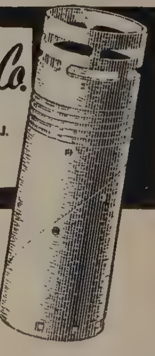
Cleveland Container's large production facilities further ensure minimum production costs as well as dependable deliveries.

Why pay more? . . . for the BEST call CLEVELAND.

* Trade Marks

Write today for our new descriptive brochure. Also ask for quotations and samples to meet your exact specifications.

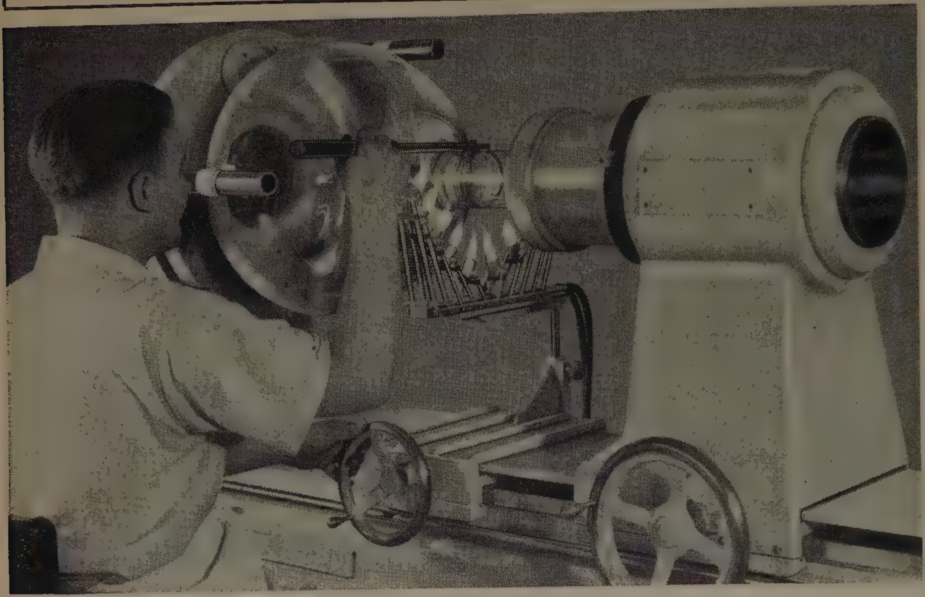
**The CLEVELAND CONTAINER Co.**
**6201 BARBERTON AVE. CLEVELAND 2, OHIO**
PLANTS AND SALES OFFICES at Plymouth, Wisc., Chicago, Detroit, Ogdensburg, N.Y., Jamesburg, N.J.
ABRASIVE DIVISION at Cleveland, Ohio
CANADIAN PLANT: The Cleveland Container, Canada, Ltd., Prescott, Ontario



*REPRESENTATIVES*

NEW YORK AREA R. T. MURRAY, 604 CENTRAL AVE., EAST ORANGE, N. J.
NEW ENGLAND R. S. PETTIGREW & CO., 10 N. MAIN ST., WEST HARTFORD, CONN.
CHICAGO AREA PLASTIC TUBING SALES, 5215 N. RAVENSWOOD AVE., CHICAGO

# LITTON INDUSTRIES NEWS



## LITTON GLASSWORKING LATHES SPEED PRECISION ASSEMBLY OF TV KINESCOPE, VACUUM TUBES

Modern vacuum tubes have extremely close alignment tolerances. Often, subassemblies must be separately aligned before junction. During sealing, both assemblies must be rotated in perfect phase to maintain this alignment.

Versatile, adaptable Litton Glassworking Lathes meet these requirements. They are built on a normalized cast iron bed, with precision ground ways and axial alignments of highest accuracy and positive phase. The lathes will chuck and hold units such as copper anodes to runouts of .001″.

Close-up of spindle head, Litton Model K lathe, showing exceptionally large diameter opening of universal chuck

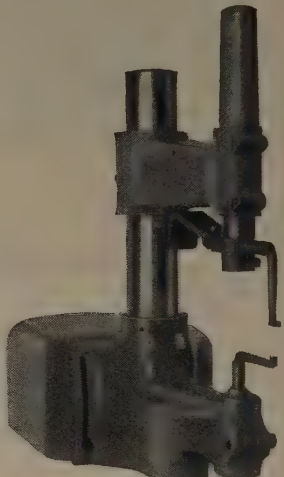
Air passages are arranged to avoid contamination, yet permit use of neutral gasses when sealing glass to metal. Burners provide the narrowest possible heating area commensurate with ample total heat. Continuously variable spindle speed, which makes possible much glassworking without blowing, is optional on all models. Foot pedals control the air or neutral gas supply, and the oxygen and gas volume. Convenient hand controls govern carburetion and air intake to the spindles.

Leading TV tube makers use Litton Glassworking Lathes to speed production of kinescope tubes 10″ to 27″ in diameter. Manufacturers find that the speed and handling ease of Litton lathes enable glassworkers to seal tube funnels to domes in minimum time—with complete control of glass distribution. Since most manufacturers align sub-assemblies on the lathe, the accurate phasing of Litton spindle heads is also an important factor.

Reliable Litton Glassworking Lathes are adaptable to the widest possible variety of glassworking jobs. Five models offer a choice of radial clearance ranging from 4″ to 17½″, and axial working lengths from 20¾″ to 75½″.

## LITTON SPOTWELDERS OFFER HIGH POWER, EXTREME FLEXIBILITY FOR PRECISION JOBS

Litton Model A Precision Spotwelder offers broad applicability of use in the manufacture of vacuum tubes. It makes possible the quick altering of production setups. Rated 2 kva continuous duty, it efficiently handles average sized or very precise jobs. Accurate alignment and absence of side play permit butt welding or parallel welding of small wires without rolling. Foot pedals and switches control top mandrel and power supply. Model A spotwelder has 6½″ throat depth and extension jaws can be added.



## SPOTWELDER TIMER

A new timer for the Litton Model A Spotwelder has been developed by Litton Industries and will be available for delivery soon. The timer employs two simple controls. One adjusts weld time in steps of 1, 2, 3, 5, 7, 10, 15, 25 and 60 cps. The other adjusts heat control in 6 steps. Proper adjustment of these controls makes possible precision welds up to the 2 kva rating of the welder.

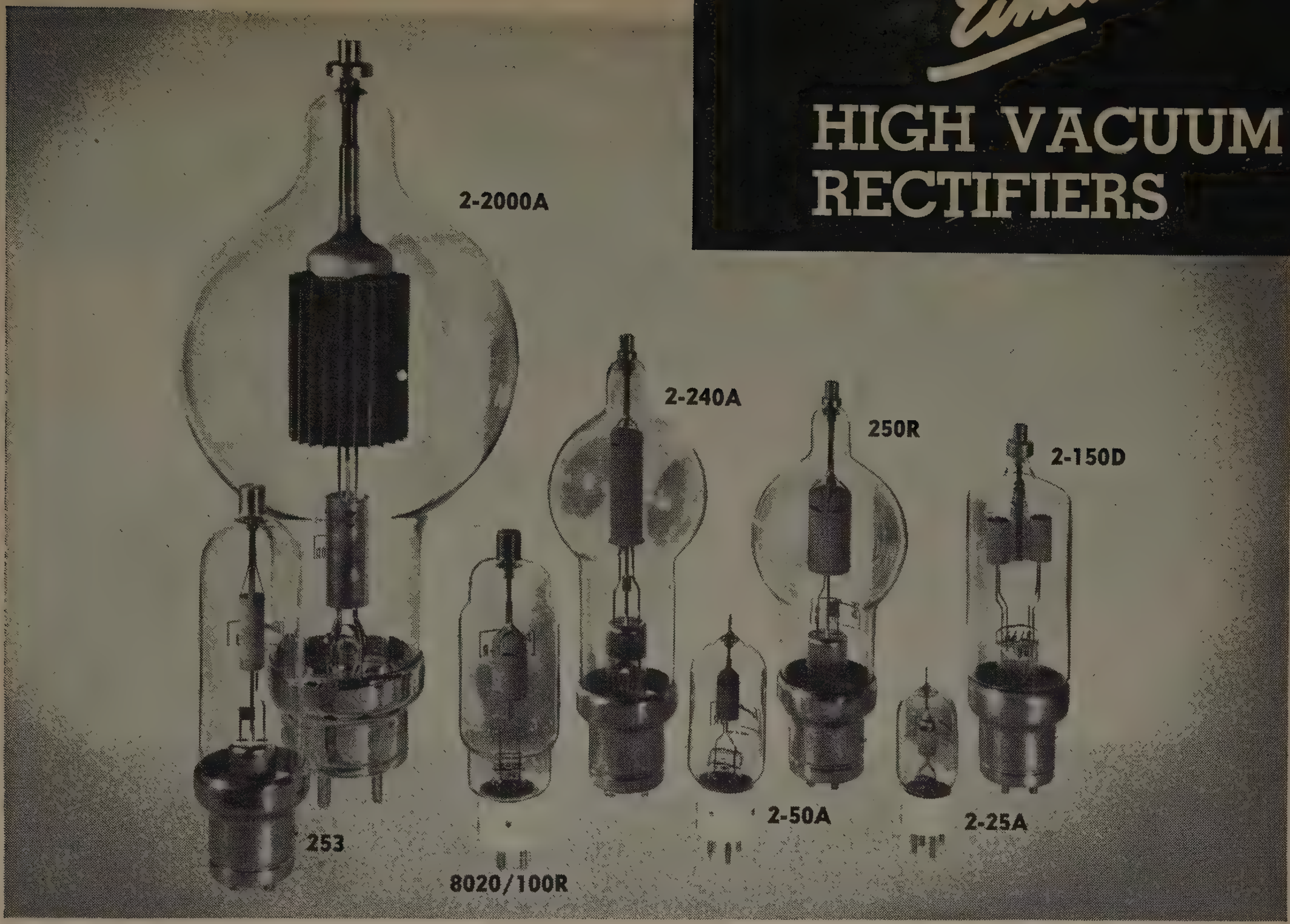
## LITTON INDUSTRIES

SAN CARLOS, CALIFORNIA, U.S.A.

**DESIGNERS AND MANUFACTURERS of:**
Glassworking Lathes and Accessories, Vertical Sealing Machines, Burner Equipment, Precision Spotwelders, Oil Vapor Vacuum Pumps, Glass Baking Ovens, Vacuum Tubes and Tube Components, Magnetrons, High Vacuum Molube Oil, Microwave Equipment.

**Eimac**

# HIGH VACUUM RECTIFIERS



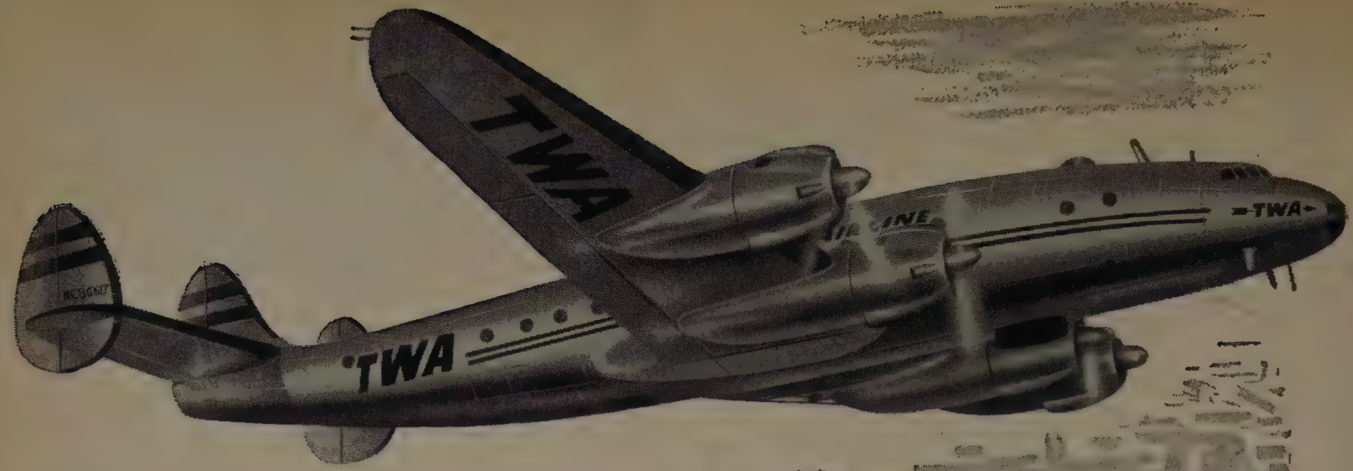
Eimac's comprehensive series of vacuum rectifiers permits a choice of "per-tube" ratings of d-c plate current from 50 ma. to 750 ma. and a choice of inverse voltage ratings from 15,000 to 75,000 volts.

These are ruggedly constructed tubes built to withstand more than normal abuse in rectifying and voltage multiplying circuits or as diode clippers. Their design incorporates many of the features long associated with the famous Eimac transmitting tubes . . . Pyrovac plates . . . thoriated tungsten filaments . . . no troublesome internal insulating materials . . . and, of course, a "hard" vacuum.

Put Eimac high vacuum rectifiers to work for you. Write today for detailed data sheets giving complete operating information and application notes.

**EITEL-McCULLOUGH, INC.**
**San Bruno, California**
Export Agents: Frazar & Hansen, 301 Clay St., San Francisco, California

| TYPE | DESCRIPTION | MAXIMUM DIMENSIONS Length Inches | MAXIMUM DIMENSIONS Diameter Inches | AVERAGE PLATE CUR. Ma. | PLATE DISSIPATION Watts | PEAK INVERSE VOLTAGE Volts | FILAMENT Volts | FILAMENT Amps. |
|---|---|---|---|---|---|---|---|---|
| **2-25A** | High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 4.5 | 1.5 | 50 | 15 | 25,000 | 6.3 | 3.0 |
| **2-50A** | High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 5.75 | 2 | 75 | 30 | 30,000 | 5.0 | 4.0 |
| **2-150D** | High vacuum rectifier. High voltage medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 8.88 | 2.75 | 150 | 90 | 30,000 | 5.0 | 13.0 |
| **2-240A** | High vacuum rectifier. High voltage, high current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 11.25 | 4 | 500 | 150 | 40,000 | 7.5 | 12.0 |
| **2-2000A** | High vacuum rectifier. High voltage, high current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 18 | 8.25 | 750 | 1,200 | 75,000 | 10.0 | 25.0 |
| **250R** | High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 10.25 | 4 | 250 | 150 | 60,000 | 5.0 | 10.5 |
| **253** | High vacuum rectifier. High current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 9 | 2.75 | 350 | 100 | 15,000 | 5.0 | 10.0 |
| **8020/100R** | High vacuum rectifier. High voltage, medium current. Instant heating, thoriated tungsten filament. Radiation cooled pyrovac plate. | 8 | 2.38 | 100 | 60 | 40,000 | 5.0 | 6.5 |



# WILCOX...

## CHOICE OF TWA

### WILCOX TYPE 96D TRANSMITTERS CHOSEN FOR *TWA's* GROUND STATION INSTALLATION

### AT ROME

TWA's far-flung empire requires the finest communications network that research and precision manufacturing can produce. TWA has long been a user of Wilcox equipment, and Wilcox is proud to be chosen again as TWA expands and improves its world-wide services. The Wilcox Type 96D 2500 watt transmitter covers the 2-26 mc. range.

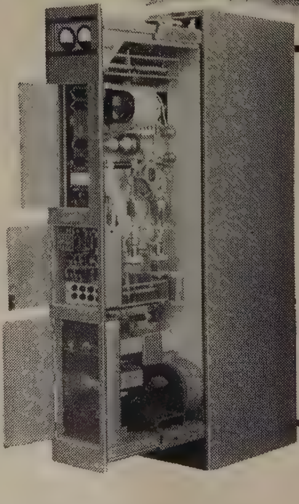
**UNIT CONSTRUCTION OFFERS OPERATING FLEXIBILITY**

Either simultaneous transmission on several frequencies or the selection of an individual frequency best suited to your particular problem.

**JAN COMPONENT STANDARDS FOR DEPENDABILITY**

All components meet the latest JAN specifications. This means maximum resistance to wear, corrosion, humidity, fungus, temperature, and time.

***Write Today*** for complete information and specifications on the Wilcox 96D Transmitting Station.



**WILCOX**
**ELECTRIC COMPANY**
Fourteenth & Chestnut
KANSAS CITY 1, MISSOURI, U.S.A.

# Better Circuit Reliability Without OPERATOR ADJUSTMENT

Sync Generator set up with compact GPL Control Console

POWER SUPPLY Built in

## the New GPL SYNCHRONIZING GENERATOR

**SMALLER THAN EXISTING UNITS**

●

**EASY MAINTENANCE**

●

**BUILT-IN POWER SUPPLY**

●

**STANDARD RELAY PANELS**

***Easy to Rack Mount***

The GPL Synchronizing Pulse Generator provides circuit reliability superior to that of comparable studio equipment. Operator adjustments are now eliminated by means of advanced circuit design, including binary counting circuits, delay-line-controlled pulse width – all operating from a stable master oscillator. The generator provides standard RTMA outputs with automatic termination of unused outputs. The AFC circuit is readily set to operate at mid-range when locked to the line.

Since the unit is smaller than existing equipment, even with its self-contained power supply, it is ideal for field operation. Swing-down panels simplify maintenance. Components are mounted on standard relay panels, facilitating studio rack mounting.

Typical of other GPL developments, the Synchronizing Generator is designed for maximum quality, operating efficiency, and dependability. Write for literature and operating information.

***Write, Wire or Phone for Details***

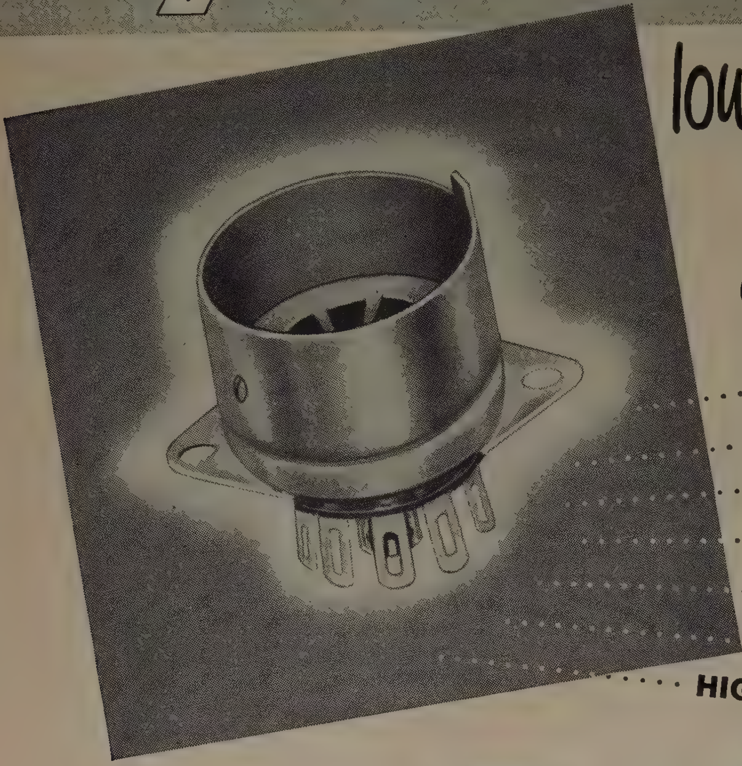
**General Precision Laboratory**

INCORPORATED

PLEASANTVILLE NEW YORK

TV Camera Chains • TV Film Chains • TV Field and Studio Equipment • Theatre TV Equipment

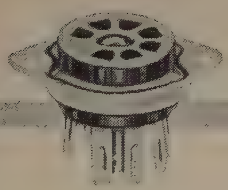
# MYCALEX

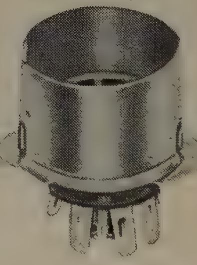


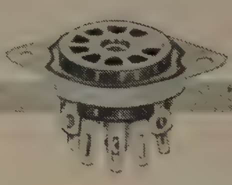
## low loss miniature TUBE SOCKETS

### OFFER ALL THESE ADVANTAGES:

- CLOSER TOLERANCES
- LOWER DIELECTRIC LOSS
- HIGH ARC RESISTANCE
- HIGH DIELECTRIC STRENGTH
- GREAT DIMENSIONAL STABILITY
- IMMUNITY TO HUMIDITY
- HIGH SAFE OPERATING TEMPERATURE







## -cost no more than PHENOLIC TYPES

These glass-bonded mica sockets are produced by an exclusive MYCALEX process that reduces their cost to the level of phenolic sockets. Electrical characteristics are far superior to phenolics while dimensional accuracy and uniformity exceed that of ceramic types.

MYCALEX miniature tube sockets, available in 7-pin and 9-pin types, are injection molded with great precision and fully meet RTMA standards. They are produced in two grades, described as follows, to meet diversified requirements.

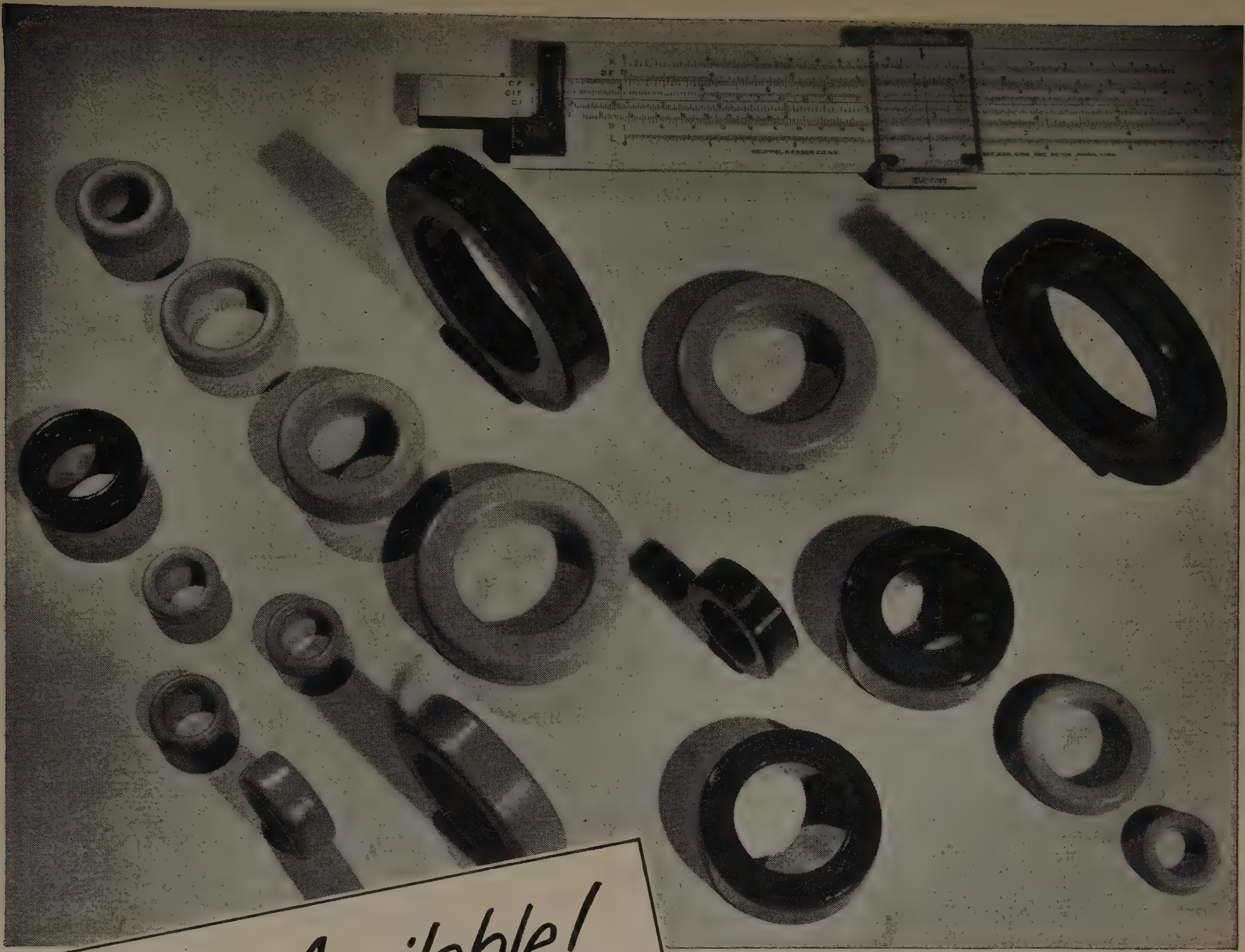
MYCALEX 410 is priced comparable to mica-filled phenolics. Loss factor is only .015 at 1 mc., insulation resistance 10,000 megohms. Conforms fully to Grade L-4B under N.M.E.S. JAN-1-10 "Insulating Materials Ceramic, Radio, Class L."

MYCALEX 410X is low in cost but insulating properties greatly exceed those of ordinary materials. Loss factor is only one-fourth that of phenolics (.083 at 1 mc.) but cost is the same. Insulation resistance 10,000 megohms.

**MYCALEX TUBE SOCKET CORPORATION**
*Under Exclusive License of*
*MYCALEX CORPORATION OF AMERICA*
30 ROCKEFELLER PLAZA, NEW YORK 20, N.Y.

SINCE 1919
MYCALEX
THE INSULATOR
TRADE MARK REG. U.S. PAT. OFF.

**CORPORATION OF AMERICA**
*"Owners of 'MYCALEX' Patents"*
Executive Offices: 30 Rockefeller Plaza, New York 20 • Plant and General Offices: Clifton, New Jersey



## *Now Available!*
## MOLYBDENUM PERMALLOY POWDER CORES*

**HIGH Q TOROIDS for use in Loading Coils, Filters, Broadband Carrier Systems and Networks—for frequencies up to 200 KC**

### COMPLETE LINE OF CORES TO MEET YOUR NEEDS

★ Furnished in four standard permeabilities—125, 60, 26 and 14.

★ Available in a wide range of sizes to obtain nominal inductances as high as 281 mh/1000 turns.

★ These toroidal cores are given various types of enamel and varnish finishes, some of which permit winding with heavy Formex insulated wire without supplementary insulation over the core.

For high Q in a small volume, characterized by low eddy current and hysteresis losses, ARNOLD Moly Permalloy Powder Toroidal Cores are commercially available to meet high standards of physical and electrical requirements. They provide constant permeability over a wide range of flux density. The 125 Mu cores are recommended for use up to 15 kc, 60 Mu at 10 to 50 kc, 26 Mu at 30 to 75 kc, and 14 Mu at 50 to 200 kc. Many of these cores may be furnished stabilized to provide constant permeability (±0.1%) over a specific temperature range.

** Manufactured under licensing arrangements with Western Electric Company.*

W&D 2930

**THE ARNOLD ENGINEERING COMPANY**

SUBSIDIARY OF ALLEGHENY LUDLUM STEEL CORPORATION

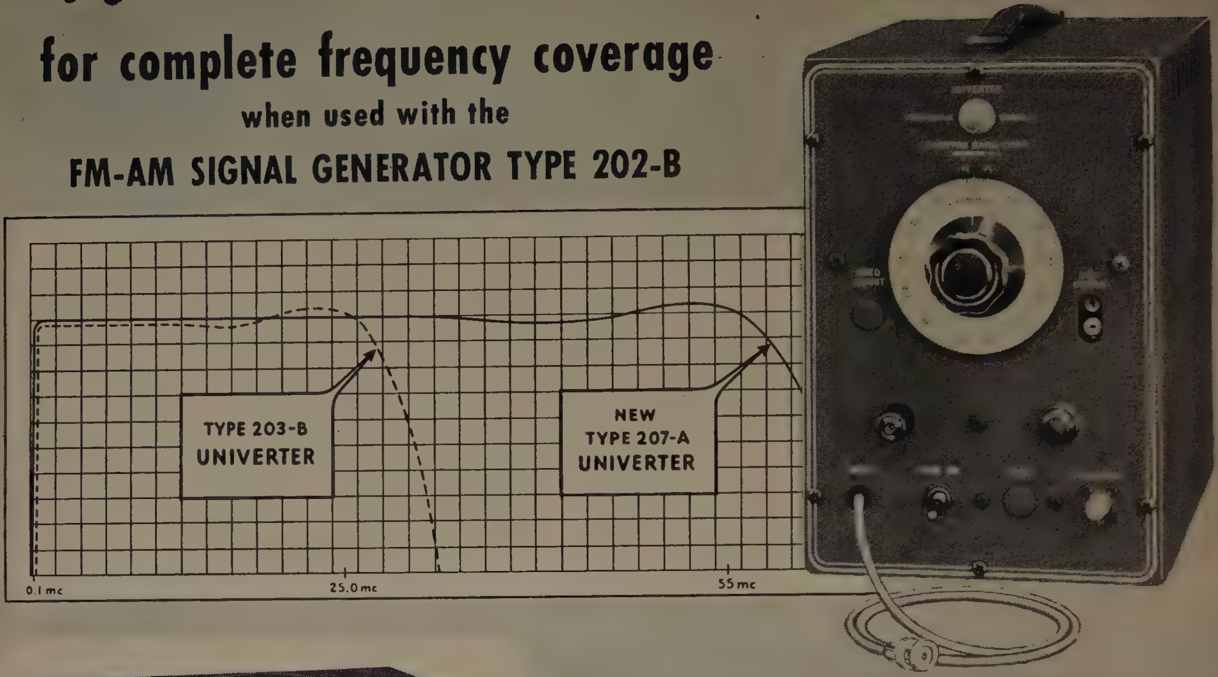
General Office & Plant: Marengo, Illinois

# *Wide Band* UNIVERTER

## for complete frequency coverage

when used with the

## FM-AM SIGNAL GENERATOR TYPE 202-B



## FM-AM SIGNAL GENERATOR

*TYPE 202-B*

- The standard signal source for the FM and TV industry.
- Univerter 207-A extends frequency range down to 0.1 mc. without change in signal level or modulation characteristics below.

**SPECIFICATIONS:**

**RF RANGES:** 54-108, 108-216 mc.
**FREQUENCY DEVIATION:** 0-24 kc., 0-80 kc., 0-240 kc.
**FM DISTORTION:** Less than 2% at 75 kc. deviation
**AMPLITUDE MODULATION:** Continuously variable 0-50%.
**RF OUTPUT VOLTAGE:** 0.1 microvolt to 0.2 volt.

## UNIVERTER

*TYPE 207-A*

The Univerter Type 207-A provides a continuous extention of the frequency range of the 202-B FM-AM Signal Generator down to 0.1 mc. The two instruments may be used over a continuous frequency range of 0.1 mc. to 216 mc. The Univerter Type 207-A subtracts 150 mc. from a signal obtained from the 202-B and provides outputs between 0.1 mc. and 55 mc. without change of signal level. Negligible spurious signals are introduced and modulation of the signal is unaffected. Small incremental changes can be made in frequency to allow the study of band pass characteristics of very narrow band receivers. A regulated power supply prevents change of gain or frequency with line voltage.

**SPECIFICATIONS** (*When used with 202-B*)

**FREQUENCY RANGE:** 0.1 mc. to 55 mc. (0.3 mc. to 55 mc. with 200 kc. carrier deviation).

**FREQUENCY INCREMENT DIAL:** Plus or minus 300 kc. calibrated in 5 kc. increments.

**FREQUENCY RESPONSE:** Flat within ±1 db over frequency range.

**FREQUENCY ADJUST:** Front panel control allows calibration with 202-B output.

**OUTPUT:** Continuously variable, at XI jack from 0.1 microvolt to 0.1 volt across 53 ohms by use of 202-B attenuator.

**HIGH OUTPUT:** Uncalibrated approximately 1.5 volts from 330 ohms into open circuit.

**DISTORTION:** No appreciable FM distortion at any level.
No appreciable AM distortion at carrier levels below 0.05 volt and modulation of 50%.

**SPURIOUS RF OUTPUT:** At least 30 db down at input levels less than 0.05 volts.

*Write for complete information*
*(In Canada, direct inquiries to R.C.A. Victor Co., Ltd., Montreal)*

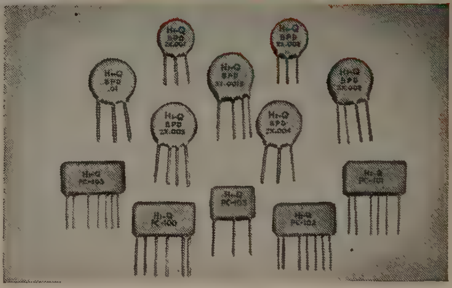
**BOONTON RADIO** *Corporation*
BOONTON, N. J.

Home entertainment has changed

. . . and in producing better and better TV
**MODERN ELECTRONICS LOOK TO HI-Q***
**Capacitors • Trimmers • Choke Coils • Wire Wound Resistors**

The fast development of the television industry since World War II has been matched, stride for stride, by **HI-Q**. For TV producers were quick to recognize this organization as their most dependable source for the ceramic components they needed in such profusion. They quickly learned that **HI-Q** engineers were competent and resourceful in developing new components to meet new needs as they arose.

Now, though **HI-Q** output has reached several million capacitors, trimmers, choke coils and wire wound resistors each month, never once have the original precision standards or strict adherence to specifications and tolerances been shaded — or the rigid system of inspection of each individual unit at each stage of production been relaxed. The **HI-Q** engineering staff is just as ready as ever to cooperate with your engineers in the production of special components for special requirements.



**HI-Q DISKS AND PLATES**

High dielectric by-pass, blocking or coupling capacitors for use where their geometrical shape makes them more adaptable than tubular components. Essentially similar, other than shape, except that in multiple units, **HI-Q** Plates do NOT have to have a common ground, as is the case with the Disk type.

**BETTER 4 WAYS**

- ✓ PRECISION
- ✓ UNIFORMITY
- ✓ DEPENDABILITY
- ✓ MINIATURIZATION

JOBBERS—ADDRESS: 740 Belleville Ave., New Bedford, Mass.

* *Trade Mark Registered, U. S. Patent Office*

**HI-Q***

*Electrical Reactance Corp.*
OLEAN, N. Y.

**SALES OFFICES:** New York, Philadelphia, Detroit, Chicago, Los Angeles

**PLANTS:** Olean, N. Y., Franklinville, N. Y. Jessup, Pa., Myrtle Beach, S. C.

# *Unique!*

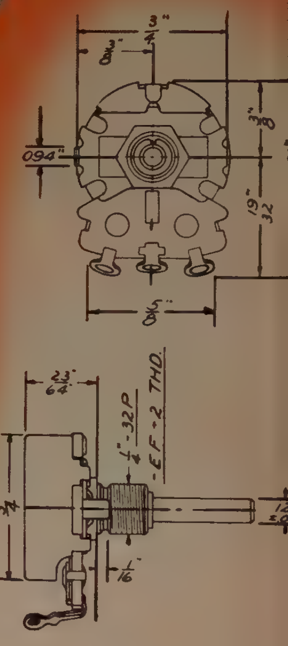
# MINIATURIZED

## VARIABLE RESISTOR for MILITARY USE

Manufactured from specially developed materials, Type 65 variable composition resistor is a unique, successful combination of miniature size and unprecedented high temperature and humidity stability. For example, Type 65 delivers top performance in fast military airplanes flying in tropical areas where both temperature and humidity vary tremendously from ground level to extreme altitudes.

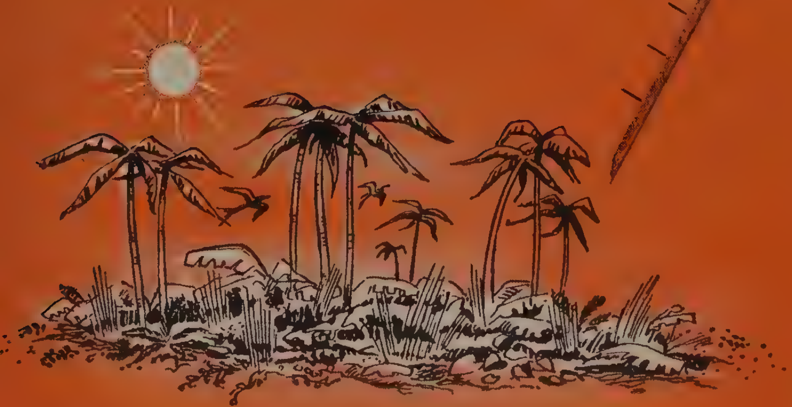
Exceptionally good delivery cycle due to tremendous precision mass production facilities.

*Please give complete details on your requirements when writing or phoning for further information.*



1/4"-32P. -
E.F.-2 THD.
3/4" DIA.
DIA. .125" ±.00

**TYPE 65**
**Photo Is Actual Size**



VARIABLE RESISTORS (COMPOSITION and WIRE WOUND)
WITH or WITHOUT ASSOCIATED SWITCH COMBINATIONS

*Specialists in Precision Mass Production of Variable Resistors*

## CHICAGO TELEPHONE SUPPL
## *Corporation*

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Phone: Walnut 2-5369

**W. S. Harmon Company**
4057 McClung Drive
Los Angeles 8, California
Phone: Axminister 3-6219

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**Jose Luis Ponte**
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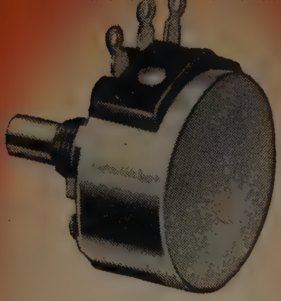
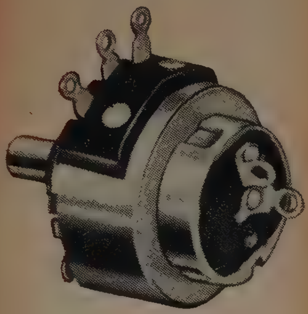

JAN Type RA 20A
2 Watt (CTS Type 252)


JAN Type RA 20B
2 Watt (CTS Type GC-252)

JAN TYPE RA 25A
3 Watt (CTS Type 25)

# *Meets* JAN-R-19

## VARIABLE WIRE WOUND RESISTORS TO ALL SPECIFICATIONS IN JAN-R-19

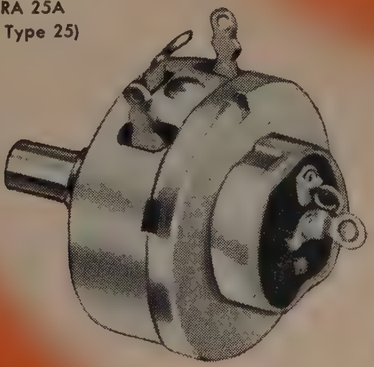
### TYPES NOW AVAILABLE

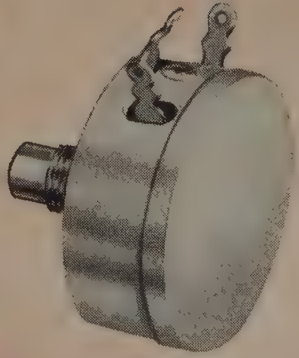
| JAN TYPE | CTS TYPE | WATT RATING |
|---|---|---|
| **RA 20A** | 252 (without switch) | 2 |
| **RA 20B** | GC-252 (with switch) | 2 |
| **RA 25A** | 25 (without switch) | 3 |
| **RA 25B** | GC-25 (with switch) | 3 |
| **RA 30A** | 25 (without switch) | 4 |
| **RA 30B** | GC-25 (with switch) | 4 |

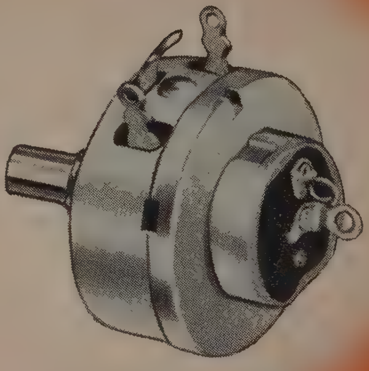
Exceptionally good delivery cycle due to tremendous precision mass production facilities.

*Please give complete details on your requirements when writing or phoning for further information.*

ILLUSTRATIONS ARE ACTUAL SIZE.


JAN Type RA 25B
3 Watt (CTS Type GC-25)


JAN Type RA 30A
4 Watt (CTS Type 25)


JAN Type RA 30B
4 Watt (CTS Type GC-25)

CTS

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*Corporation*

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hone: Walnut 2-5369

V. S. Harmon Company
638 South La Cienega Blvd.
os Angeles, California
hone: Axminister 3-6219

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**OTHER EXPORT**

Sylvan Ginsbury
8 West 40th Street

A COMPLETE LINE OF VARIABLE RESISTORS (COMPOSITION AND WIRE WOUND) FOR MILITARY APPLICATIONS.

*Specialists in Precision Mass Production of Variable Resistors*

FOUNDED 1896

**TRUSCON**...*a name you can build on*

## Truscon Builds

# World's Tallest Radio Tower

Truscon has fabricated for the United States Government a 1,212-foot tower recently erected near Rome, New York.

The 1,212-foot giant overshadows every other man-made structure in the world, except the 1,250-foot high Empire State building. It is truly a tribute to the skill of the men at Truscon who designed, engineered, and produced it. Not many years ago a tower of this type and height would have been considered impractical to build. Work on the structure began on Truscon's drawing boards in the spring of 1948. Erection was started in September 1950.

Nearing completion last November, the tower had its first test of consequence during the storm that brought record snow and 125-mile-an-hour winds to the eastern section of the country. In this blow the tower swayed approximately seven-tenths of the seven feet it is calculated to sway in a 150-mile-an-hour hurricane.

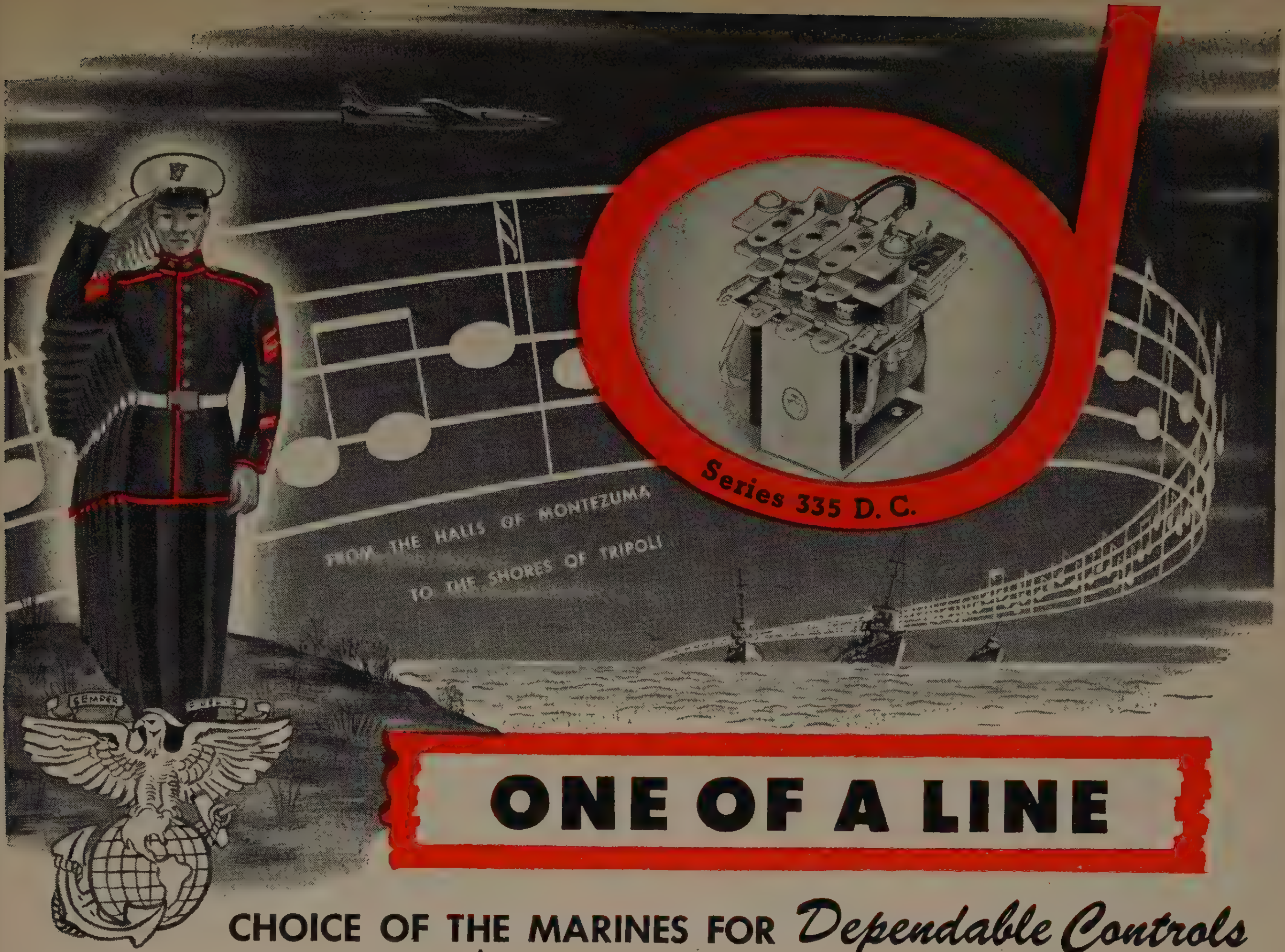
Requiring 772 tons of fabricated steel, the great structure is supported by 4 miles of guy cables, most of which are anchored almost a quarter of a mile away from the base. The new tower will be used by scientists at the Griffiss Air Force Base near Rome (N. Y.) for the study of Loran, a radio navigation aid first developed during the last war.

Today, rising skyward in many nations are many hundreds of Truscon-built "fingers of steel" over which pour communications for the attentive ears and eyes of the world's people. Your phone call or letter to any convenient Truscon district office, or to our home office in Youngstown, will bring you immediate, capable engineering assistance on your tower problems. Call or write today.

*A construction photograph shows giant tower's base. More than 1400 cu. yds. of concrete was used to form base pier and guy anchors.*

TRUSCON
MARK OF MERIT PRODUCTS

**TRUSCON® STEEL COMPANY**
Subsidiary of Republic Steel Corporation
Youngstown 1, Ohio


Series 335 D.C.
FROM THE HALLS OF MONTEZUMA
TO THE SHORES OF TRIPOLI
SEMPER FIDELIS

# ONE OF A LINE

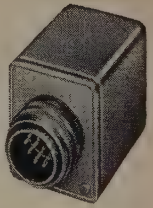
## CHOICE OF THE MARINES FOR *Dependable Controls*

**GUARDIAN**

**RELAYS**

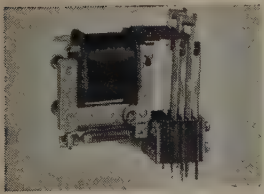
"From the halls of Montezuma" to marine testing laboratories—technicians are singing the praise of Guardian Relays. The Series 335 D.C. Relay shown above is but one of a most comprehensive line of Guardian Relays *approved* by the Marine Corps and all branches of the Armed Forces. Available with open-type construction or *hermetically sealed*—Guardian's Series 335 D.C. Relay or any Government Approved Guardian control unit used singly, in combination, or in completely packaged controls, will serve you faithfully and well.

*The Series 335 D.C. Guardian Relay shown above is available with the A.N. Connector Plug, Octal Plug and Lug Header* hermetic seal *containers. Packs loads of power over a wide operating range, withstands the rigors of dust, moisture, salt air, temperature changes, vibration and impact.*

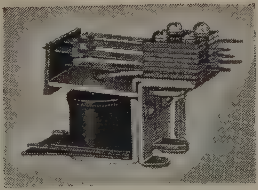


A.N. CONNECTOR PLUG
HERMETIC SEALED
CONTAINER

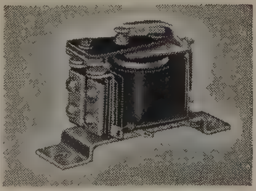
**NEW CATALOG** on Hermetically Sealed Guardian Relays with various containers is yours for the asking, cost-free.



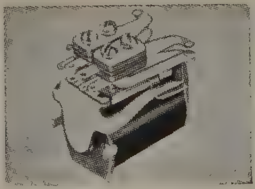
Series 30 A.C.



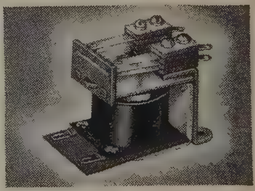
Series 210 A.C.—215 D.C.



Series 220 A.C.



Series 595 D.C.



Series 610 A.C.—615 D.C.

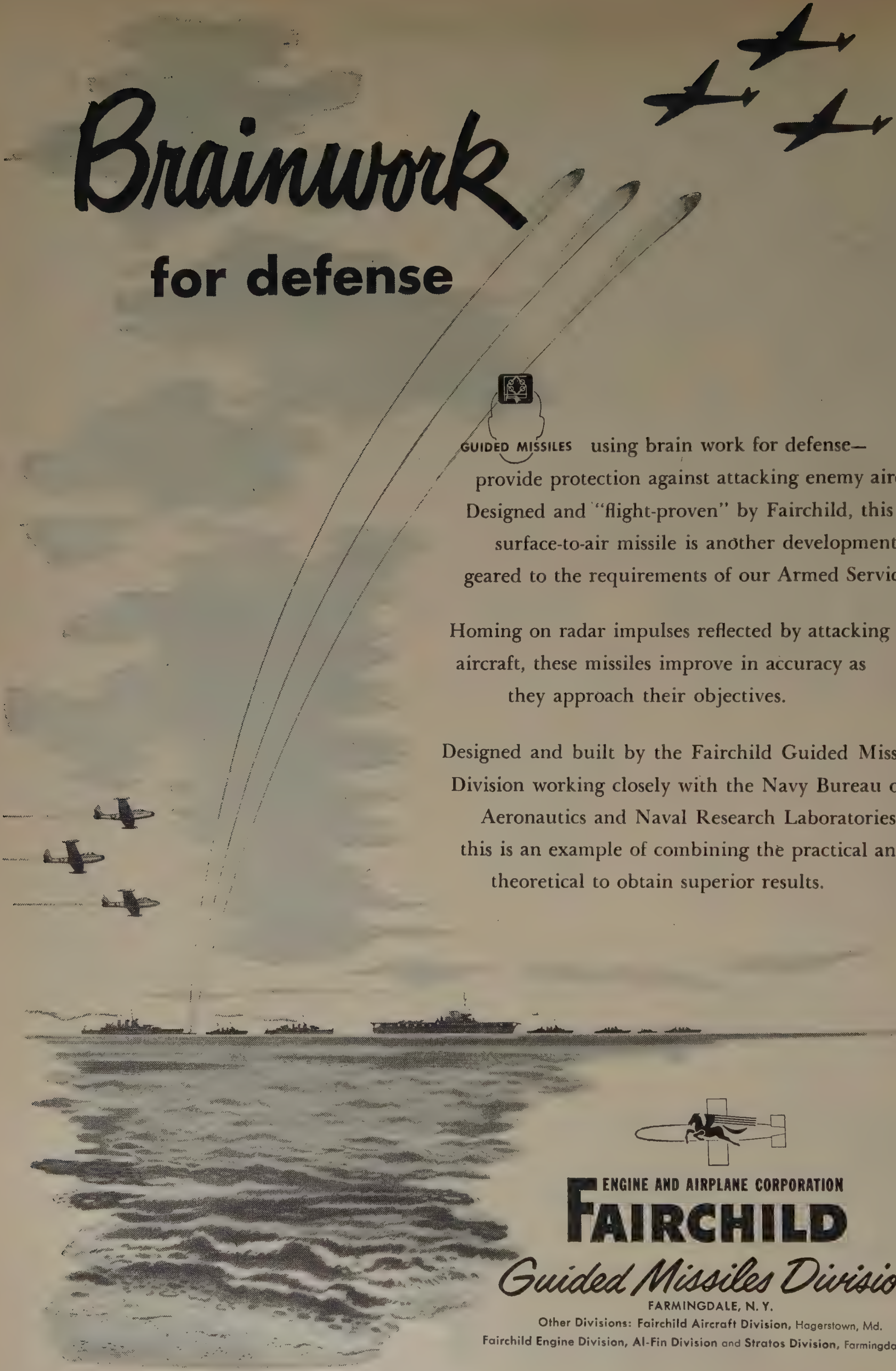
*WRITE OR WIRE . . . FREE CATALOG, SPECIFIC RECOMMENDATIONS, NO OBLIGATION.*

**GUARDIAN ELECTRIC**

1628-G W. WALNUT STREET
CHICAGO 12, ILLINOIS

A COMPLETE LINE OF RELAYS SERVING AMERICAN INDUSTRY

# Brainwork
## for defense



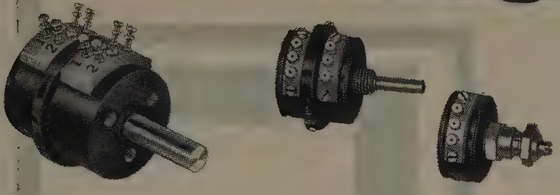
**GUIDED MISSILES** using brain work for defense—provide protection against attacking enemy aircraft. Designed and "flight-proven" by Fairchild, this surface-to-air missile is another development geared to the requirements of our Armed Services.

Homing on radar impulses reflected by attacking aircraft, these missiles improve in accuracy as they approach their objectives.

Designed and built by the Fairchild Guided Missiles Division working closely with the Navy Bureau of Aeronautics and Naval Research Laboratories, this is an example of combining the practical and theoretical to obtain superior results.

ENGINE AND AIRPLANE CORPORATION
**FAIRCHILD**
*Guided Missiles Division*
FARMINGDALE, N. Y.
Other Divisions: Fairchild Aircraft Division, Hagerstown, Md.
Fairchild Engine Division, Al-Fin Division and Stratos Division, Farmingdale, N. Y.

# There's a TIC® Potentiometer for every application



**RVP3* High Precision machined aluminum base Potentiometers** . . . available in models for either linear or non-linear functions with standard resistance values up to 200,000Ω. Linearity to ±0.1%. Eleven gang assembly shown — example of TIC's potentiometers multi-ganged with TIC's adjustable clamp ring. Can be supplied to meet various mounting requirements — single hole, 3 tapped hole mounting or servo mounting as desired.

**Miniaturization of precision potentiometers** is keeping pace with the increased demand for smaller assemblies and compact design. Now you can minimize wasted space with TIC's outstanding, new RV⅞* and RV1-1/16* Miniature Potentiometers.

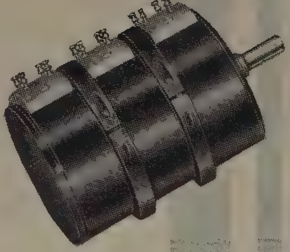
In spite of their thumbnail size the RV⅞ and the RV1-1/16 are precision, high linearity variable resistors, (or adjustable trimmers) of high stability — achieving a standard of performance hitherto unavailable in such miniature potentiometers.

Construction features include: precision machined aluminum base . . . low torque . . . all soldered connections except sliding contacts . . . paliney contacts . . . can be sealed to withstand all humidity, salt spray and altitude specifications. Ganging if desired with TIC adjustable clamp ring.

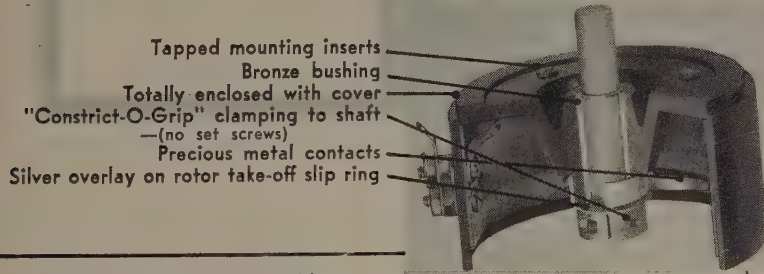
RV⅞* available with linear resistance elements *only* — nine standard resistance values from 100 to 25,000 ohms. Power rating 6 watts at 25°C. Illustration shows RV⅞ with threaded bushing . . . servo mounting available if desired.

RV1-1/16* available with linear or non-linear resistance elements — nine standard resistance values from 100 to 50,000 ohms. Illustration shows RV1-1/16 with 3 tapped hole mounting . . . servo mounting or threaded bushing if desired.

**Sine and cosine potentiometers available in RVP3* and RV2* bases.**

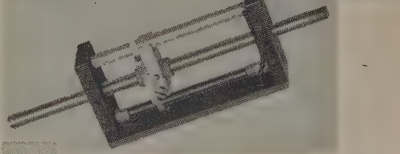


**Type RV1⅝* and RV2* High Precision Potentiometers** . . . semi-standardized types of precision machined aluminum base potentiometers with exceptionally high electrical accuracy and mechanical precision. For both linear and non-linear functions. Designed for precision instrument, computer and military applications. Accurate phasing of individual units possible with clamp-ring method of ganging. Ball bearing models available.


Tapped mounting inserts
Bronze bushing
Totally enclosed with cover
"Constrict-O-Grip" clamping to shaft —(no set screws)
Precious metal contacts
Silver overlay on rotor take-off slip ring

**Type RVT Translatory Potentiometers** . . . actuated by longitudinal instead of rotary motion providing linear electrical output proportional to shaft displacement. Used as a position indicator, high amplitude displacement type pickup and for studying low frequency motion or vibration. Features exceptionally high linearity and resolution. Available in various lengths and resistance values.

**Type RV3* Bakelite Base Precision Potentiometers** . . . available in models for either linear or non-linear functions. Stock resistance values ranging from 100Ω to 200,000Ω and power ratings of 8 and 12 watts. 360° mechanical rotation or limited by stops as desired. Potentiometers of this type available to widely varying accuracy requirements (linearity to ±0.25%) — see TIC Bulletin RV3-250. Special models available for high humidity applications.



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Telephone: Acton 600

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**Numbers refer to diameter of bases.*

# Another RCA FIRST!

# The new RCA 21″ metal rectangular kinescope featuring electrostatic focus

**Now . . .** RCA electrostatic focusing in a super-large 21″ kinescope . . . combining all the advantages of this new RCA technical achievement with the superior constructional features of metal-shell picture tubes. The tube provides television pictures of equal or better quality than magnetically focused types and is a lasting, long-range engineering accomplishment which goes beyond the present program to conserve critical materials.

The new RCA-21DP4 has a maximum over-all length of only 22⅝″, and a picture area of 18⅜″ x $13\frac{15}{16}$″. The frosted Filterglass face is made of high-quality glass, provides improved contrast, and minimizes specular reflection. Since the tube utilizes the structural strength of steel, and weighs substantially less than a comparable all-glass tube, it can be safely shipped in the receiver.

The RCA-21DP4 employs an electron gun of improved design that provides good uniformity of focus over the entire picture area. Focus can be maintained automatically with variation in line voltage and with adjustment of picture brightness. Because the focusing electrode draws very little current, the voltage for the focusing electrode can be provided easily and economically. Design-center maximum voltage rating is 18 kilovolts, diagonal-deflection angle 70°, and horizontal-deflection angle 66°.

**RCA Application Engineers** are ready to consult with you on the application of the RCA-21DP4 and its associated components to your specific designs. For further information, write RCA, Commercial Engineering, Section 47FR, Harrison, N. J., or your nearest RCA field office.

* * *

FIELD OFFICES: (EAST) Harrison 6-8000, 415 S. 5th St., Harrison, N. J. (MIDWEST) Whitehall 4-2900, 589 E. Illinois St., Chicago, Ill. (WEST) Trinity 5641, 420 S. San Pedro St., Los Angeles, California.

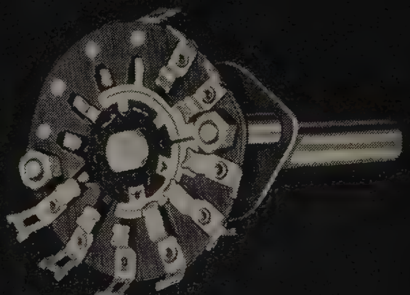
The Fountainhead of Modern Tube Development is RCA

**RADIO CORPORATION of AMERICA**

ELECTRON TUBES HARRISON, N. J.

# NEW MINIATURE SWITCHES BY CENTRALAB

STANDARD SIZE



NEW MINIATURE SIZE

*Now, Centralab, the first name in Electronic Components, offers new design possibilities in an entirely new line of truly miniature, top-quality, rotary switches for radio, TV and similar high-frequency, low-power applications.*

**WANT MORE INFORMATION?** **SEE NEXT 2 PAGES ➡**

# CENTRALAB ANNOUNCES A COMPLETE

**FOR COMMERCIAL OR**

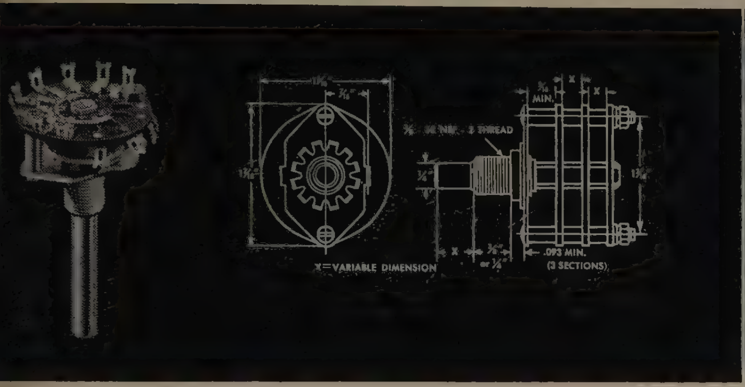
## New Centralab Miniature Switches Give You Top-Quality Construction & Performance

You'll find these new Centralab miniature rotary switches have every quality-construction feature — every quality-performance characteristic as previously built into larger size Centralab switches. Switch sections are high-quality, laminated phenolic or Centralab's own Grade L5 Steatite. Clips and contacts are self-cleaning, spring brass with heavy silver plating or coin silver. All other metal parts are adequately plated to meet commercial or government specifications. Switches have positive 30° indexing. What's more, these miniature Centralab switches offer identical opportunity with the larger size units for having multiple sections on a single shaft. Switches are available with 2 to 11 positions with stops, or 12 position continuous rotation. Of particular interest for AM-FM-TV designers — these miniature switches may be furnished with dual concentric shafts coupled with regular line switches and Centralab's famous Model 2 Radiohm variable resistors. (See illustrations at right.)

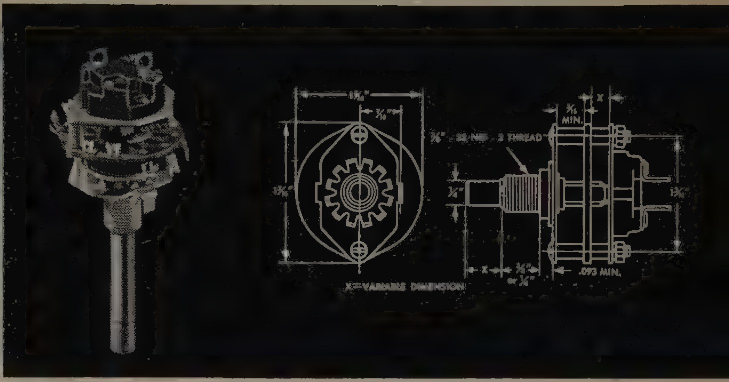
# NEW LINE OF MINIATURE SWITCHES

**MILITARY APPLICATIONS**

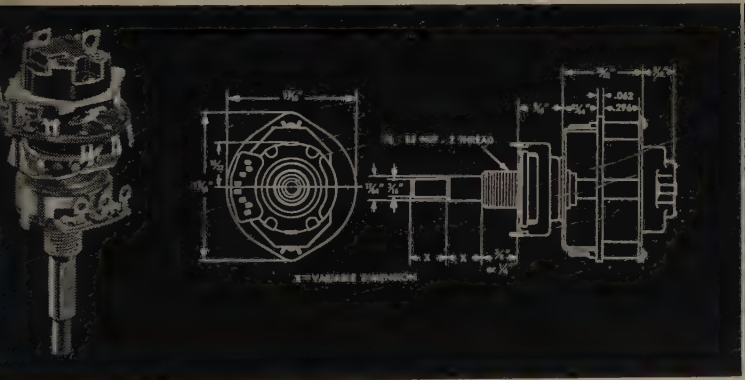
## Now Centralab offers a completely new, unusually small rotary switch line — available in a variety of multi-pole, multi-position, multi-section models and in combination with line switches and variable resistors.



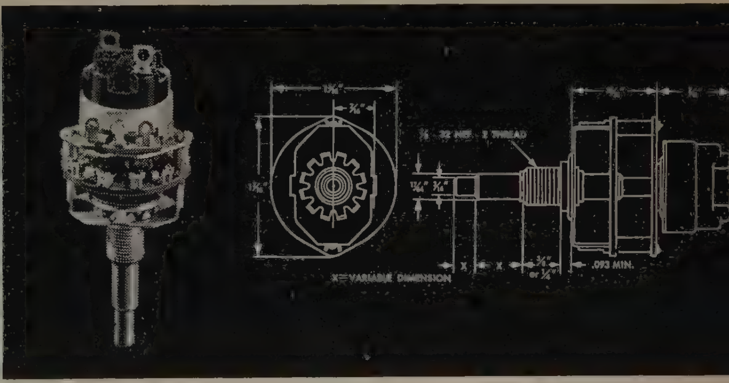
New Centralab Series 20, miniature, single section, phenolic switch. Available in 2 to 11 positions with stops, or 12 position continuous rotation — and with multiple sections.



Here's standard Series 20 miniature switch with standard shaft and phenolic section with conventional off-on switch added. Also available with multiple sections.



Combination Series 30 miniature switch unit with dual concentric shaft — permits independent operation of miniature switch including off-on switch, and Model 2 variable resistor.



Same combination Series 30 unit as shown at left, *except* that Model 2 variable resistor is mounted at rear of miniature switch. Position of resistor provides convenience of wiring.

If you need a truly small size, long-life switch, Centralab's new miniature Series 20 and Series 30 switches are the answer. They have been specifically designed to meet the modern trend toward greatly reduced size in electronic equipment for high-frequency, low-current applications. Extremely compact design and small size, plus availability of separate sections and index assemblies provide an adaptability that is invaluable to design engineers and manufacturers. For complete specification information on the new Centralab Miniature Series 20 and Series 30 Switch line, mail the coupon today. *Manufacturer's samples promptly.*

**Centralab** CRL ®

**Division of GLOBE-UNION INC • Milwaukee**

**CENTRALAB** Division of Globe-Union Inc.
920 East Keefe Avenue, Milwaukee 1, Wisconsin

Please send me complete specification information on Centralab's new Series 20 and Series 30 Miniature Switch line.

Name ..............................................................................

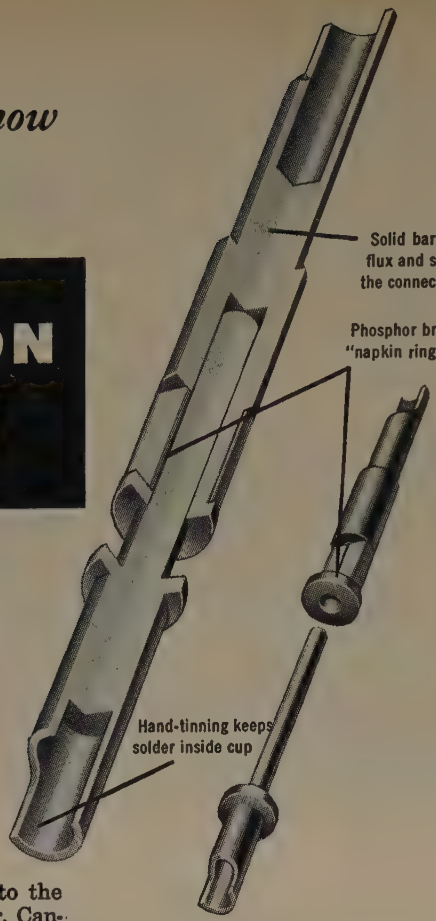
Address ..........................................................................

City .................................................... Zone ....... State ..............................

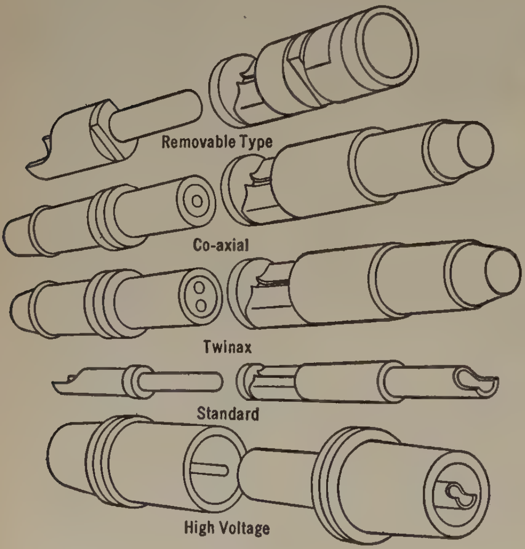
*Here's why those in the know demand–*

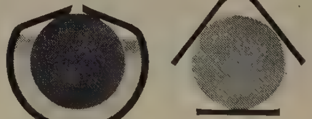
**CANNON PLUGS**

*superior contact design*


Solid bar stock keeps flux and solder out of the connector.
Phosphor bronze "napkin ring"
Hand-tinning keeps solder inside cup

Because the contact is the key to the success of any electric connector, Cannon has always applied the highest order of skill and care to this all-important detail. Cannon pin and socket contacts are all precision machined from solid bar stock. Silver or gold plating maintains high conductivity after years of constant use. Phosphor bronze "napkin ring" of the socket keeps pressure on large areas of heavy metal, preventing current loss. There are no thin metal tangent contact points in Cannon contacts. (See below). Solder cups are carefully tinned by hand to keep the solder inside the cup. Cannon socket contacts are full floating to assure perfect alignment. You'll find these design features throughout the great variety of precision contacts used in all Cannon connectors. For real value demand Cannon.





Cannon design (above left) makes contact on large, heavy metal surfaces. Current is not carried through spring section. In Cannon Connectors there are no thin metal tangent contact points, like the design shown at right.

**CANNON ELECTRIC**

*Since 1915*

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REPRESENTATIVES IN
PRINCIPAL CITIES

In Canada & British Empire: Cannon Electric Co., Ltd., Toronto 13, Ontario. World Export (Excepting British Empire): Frazar & Hansen, 301 Clay St., San Francisco, California.

## SECTION MEETINGS

### ATLANTA

"Highlights of IRE Convention held in New York," by E. S. Lammers, W. A. Edson, and M. A. Honnell; March 30, 1951.

### BALTIMORE

"Recent Developments in Microphones and Loudspeakers," by H. F. Olson, RCA Laboratories; April 11, 1951.

### BEAUMONT-PORT ARTHUR

"Electronic Reservoir Analyzer," by G. C. R. Dutton and Sinclair Balch, Sun Oil Company; April 12, 1951.

### BOSTON

"A New Coupling Circuit for High-Quality, High-Efficiency Amplifiers," by F. L. McIntosh, McIntosh Laboratory; November 28, 1950.

"Demonstration Lecture on Sound and Ultrasound," by R. H. Bolt, Faculty, Massachusetts Institute of Technology; February 27, 1951.

"Use and Interpretation of Noise and Distortion Meters," by Charles Cady, General Radio Company; March 8, 1951.

"Radio Astronomy," by H. C. Van De Hulst, Leyden University; March 29, 1951.

### BUFFALO-NIAGARA

"Synchroflection: A Horizontal Deflection System Possessing Inherent Noise Immunity," by K. R. Wendt and W. K. Squires, Sylvania Electric Products, Inc.; March 28, 1951.

### COLUMBUS

"The Fluid Mapper," by A. D. Moore, Faculty, University of Michigan; March 29, 1951.

"The Empire State TV Story," by R. W. Masters, Faculty, Ohio State University; April 11, 1951.

### DALLAS-FORT WORTH

"Effects of Electric Current on the Human Body," by J. W. Ford, "Neuro-Electronic Control," by J. W. Harris, and "The Limitation of Disk Recording," by A. B. Welch, Students, Southern Methodist University; March 19, 1951.

### DENVER

Tour of Microwave Radio Telephone and Television System, Broadcasting Network Control Facilities, and Teletypewriter Control Facilities of the American Telephone and Telegraph Company, by P. W. Harrison, American Telephone and Telegraph Company; March 13, 1951.

"High Frequency Transmission and Reception," by Harvey Wehrman, Radio Station KLZ, Denver, Colorado, and Film: "Effects of Atomic Warfare"; April 10, 1951.

### DETROIT

"Fundamentals of Analog Computation," by W. A. Wheatley, Willow Run Analog Computer Facilities; March 16, 1951.

"An Electronic Planimeter," by J. J. Casey and N. R. Brainard, "The Vectograph," by R. S. Fleming and M. E. Quinn, and "An Analogue of Static Loading," by W. E. Chapelle, Students, University of Detroit; April 20, 1951.

### EMPORIUM

"Four Levels of Interest in Technical Reading," by P. H. Wendel, Wendel Publishing Company; April 10, 1951.

### EVANSVILLE-OWENSBORO

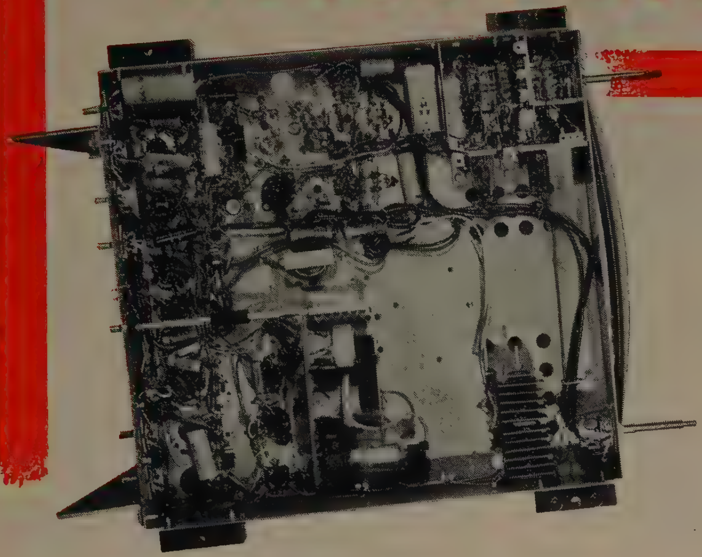
"Digital Computers," by E. J. Garvey, International Business Machines Company; March 14, 1951.

"Transcontinental Television Systems," by H. R. Gruelle, Indiana Bell Telephone Company; April 11, 1951.

*(Continued on page 38A)*

*A Good Example of*

# MODERN ENGINEERING*



# MOTOROLA'S

## NEW TS-196 TELEVISION CHASSIS

A total of 27 RMC By-pass and Temperature Compensating "DISCAPS" are incorporated in this new Motorola TV chassis. Motorola engineers specified them in place of tubular ceramic and mica condensers because of their outstanding performance and cost savings as well as their availability in these trying times.

Engineering costs out of a TV or radio chassis without sacrificing quality or performance is a mighty important job today. You, too, will find an important cost savings by specifying RMC By-pass and Temperature Compensating "DISCAPS."

If you will advise us of the types of tubular ceramic and mica condensers which you are now using, we will send samples of RMC "DISCAPS" with prices. The savings will be very apparent.

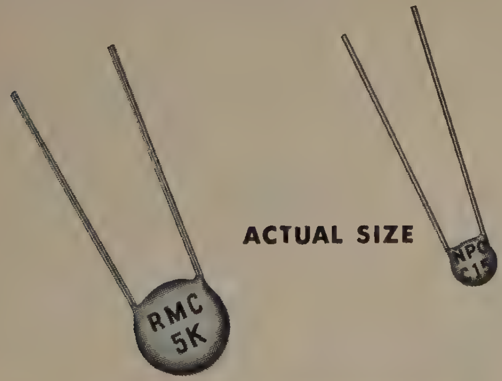
**RMC "DISCAPS"**
**THE RIGHT WAY TO SAY**
**CERAMIC CAPACITORS**

*Engineering with an Eye to Quality and Cost Savings

# RMC DISCAPS

## Used in this New Motorola Chassis

| BY-PASS TYPES | TEMPERATURE COMPENSATING TYPES |
|---|---|
| 7 – .005 MFD | 1 – 150 MMF ±20% N-750 |
| 7 – .001 MFD | 2 – 30 MMF ±10% N-150 |
| 4 – .0015 MFD | 1 – 68 MMF ±10% N-470 |
| 2 – 2×.008 MFD | 1 – 180 MMF ±20% 3000 volt |



**DISCAP**
**CERAMIC**
**CONDENSERS**

**RMC**

**RADIO MATERIALS CORPORATION**
GENERAL OFFICE: 1708 Belmont Ave., Chicago 13, Ill.
FACTORIES AT CHICAGO, ILL. AND ATTICA, IND.

*Two RMC Plants Devoted Exclusively to Ceramic Condensers*

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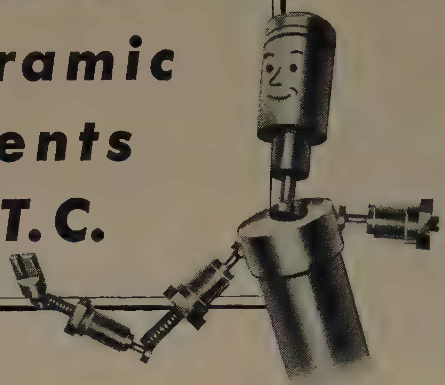
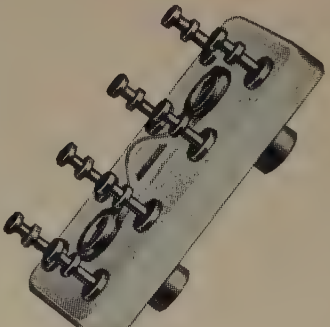
3 new ceramic components from C.T.C.



**LS-8 SLUG TUNED COIL FORM.** Features silver-plated phosphor bronze clip terminals which cannot loosen. Height, 23/32″ Maximum diameter, ½″. Mounts in "D" punched hole or in ¼″ round hole. Coil form is of grade L-5 silicone impregnated ceramic. Slug is provided with a spring lock. All metallic parts except clips are cadmium plated. Supplied complete with slug and all mounting hardware.

**X1986 CERAMIC TERMINAL BOARD.** Grade L-5 silicone impregnated ceramic. 35/64″ O.A. mounted height including terminals. 1¼″ length, ⅞″ width. Assembly has 8 C.T.C. Type X1558 terminals in 2 rows (4 per row) 9/16″ apart, plus two 4-40 tapped standoffs 3/16″ high on ⅝″ centers. A center ground strap is provided to which standoffs are riveted and soldered for good grounding at R.F. frequencies. All metal parts are non-ferrous, heavily plated.



**X1990 CERAMIC FEED-THROUGH INSULATOR.** Grade L-5 silicone impregnated ceramic. ⅞″ O.A. length including through terminal. ⅜″ hex bushing threaded for ¼″ hole mounting. Voltage breakdown 4800 volts R.M.S. @ 60 cycles AC.

**SPECIAL CONSULTING SERVICE**

A staff of thoroughly experienced C.T.C. component engineers is at your service . . at no cost to you . . . to help you secure exactly the *right* components for your requirements. They will design special units for you when standard parts are not suitable, working closely with you for economical, satisfactory results.

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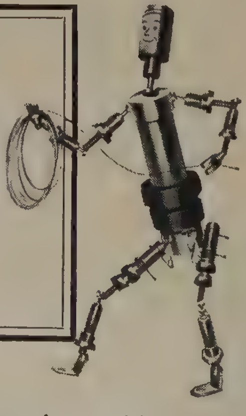


CTC

*custom or standard . . . the guaranteed components*

## SECTION MEETINGS

*(Continued from page 36A)*

FORT WAYNE

"Magnetic Amplifier Circuits, Components and Applications," by E. G. Downie, General Electric Company; March 22, 1951.

HOUSTON

"A Laboratory Class C Amplifier," by H. C. Dillingham, Faculty, Texas A & M College; March 20, 1951.

KANSAS CITY

"Capacity Operated Relays," by R. A. Kleist, Student, University of Kansas; "Color Television," by R. M. Thoman, Student, Kansas State College; and "Indicating Circuits and Equipment for Radiation Counters," by J. N. Jones, Student, University of Arkansas; April 5, 1951.

"Dialing Over Long Distance Telephone Lines," by R. F. Landry, American Telephone and Telegraph Company; April 19, 1951.

LOS ANGELES

"Impedance Measuring Techniques; 50 to 1,000 mc," by Frederick Ireland, General Radio Company, and "Engineering A Printed Circuit TV Tuner," by Hubert Shaw, Hoffman Radio Corporation; April 3, 1951.

LOUISVILLE

"Adventures in Research," by R. C. Dostal, Westinghouse Research Laboratories; March 29, 1951.

MIAMI

"Modern System and Equipment Developments for Adjacent or Split Channel Communications," by L. P. Morris, Motorola Corporation; March 30, 1951.

MONTREAL

"Recent Developments in Microphones and Loudspeakers," by H. F. Olson, Acoustical Laboratory, RCA, Princeton, New Jersey; March 7, 1951.

"Electronic Field Problems and Their Solution," by J. T. Thwaites, Canadian Westinghouse Company; March 28, 1951.

NEW MEXICO

"Telemetering Antenna," by S. S. Locus, Sandia Corporation; December 22, 1950.

"The Man Behind the Push Button," by C. J. Breitwieser, Consolidated Vultee Aircraft Corporation; February 23, 1951.

"Engineering Aspects of Digital Computers," by J. H. Parsons, Computer Group; March 30, 1951.

NEW YORK

"Current Problems of Frequency Allocation," by C. B. Plummer, Federal Communications Commission; March 7, 1951.

NORTH CAROLINA-VIRGINIA

"Design Work on North Carolina State College Nuclear Reactor," by A. C. Menius, Jr., Faculty, North Carolina State College; April 6, 1951.

OMAHA-LINCOLN

"Description and Inspection of the Facilities of Station KMTV," by Ray Schroeder, Station KMTV; March 19, 1951.

OTTAWA

"Millimicrosecond Pulse Techniques," by G. J. R. MacLusky, National Research Council Atomic Energy Project; March 29, 1951.

PHILADELPHIA

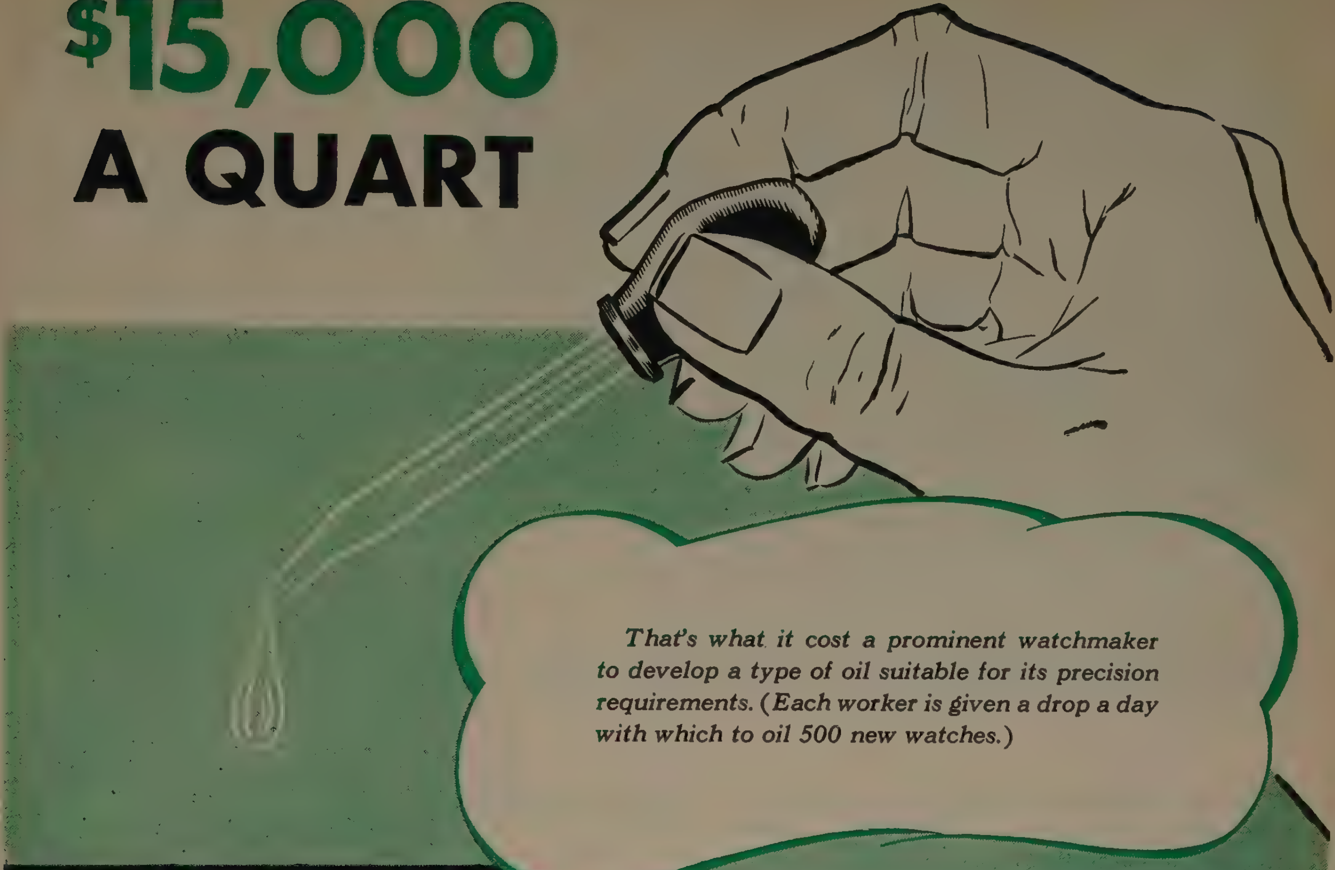
"Submarine Detection by Sonar," by J. J. Coop, Submarine Warfare Division, N.A.D.C.; April 5, 1951.

PHOENIX

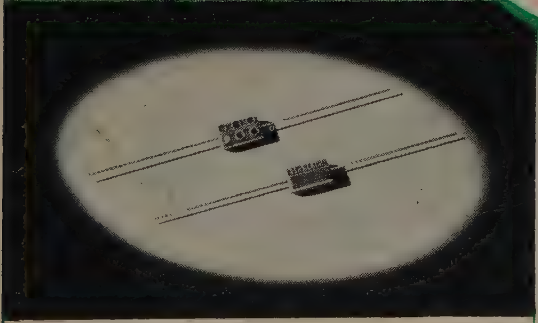
"Institute Affairs and Section Organization," by Austin Eastman, Faculty, University of Washington; March 17, 1951.

*(Continued on page 40A)*

# $15,000 A QUART



*That's what it cost a prominent watchmaker to develop a type of oil suitable for its precision requirements. (Each worker is given a drop a day with which to oil 500 new watches.)*



CM-15 MINIATURE CAPACITOR

Actual Size 9/32″ x 1/2″ x 3/16″

For Television, Radio and other Electronic Applications.

2 mmf.-420 mmf. cap. at 500v DCw.

2 mmf.-525 mmf. cap. at 300v DCw.

Temp. Co-efficient ± 50 parts per million per degree C for most capacity values.

6-dot color coded.

That quantity is no indication of quality is amply demonstrated by the CM-15 El Menco Capacitor. Tiny as it is, it will give sustained superior performance under the most adverse conditions.

Before it leaves the factory it is tested at *double* its working voltage to *insure* this unfailing performance.

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**Willimantic, Connecticut**

MANUFACTURERS ARE INVITED
TO SEND FOR SAMPLES

**MOLDED MICA El-Menco MICA TRIMMER**

**CAPACITORS**

FOREIGN RADIO AND ELECTRONIC MANUFACTURERS COMMUNICATE DIRECT WITH OUR EXPORT DEPT. AT WILLIMANTIC, CONN. FOR INFORMATION.

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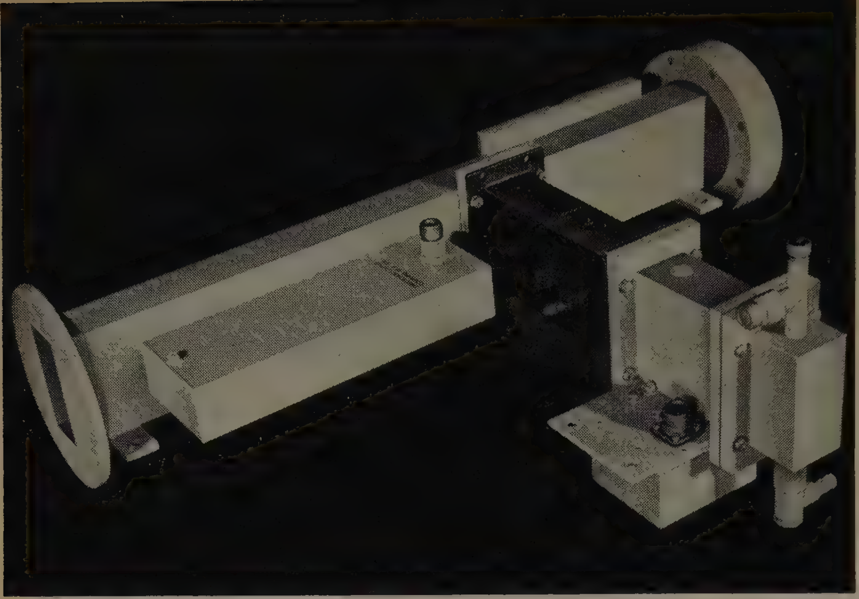

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This is another example of the type of help Terpening is set up to provide prime contractors on microwave transmission line systems—from design through production. Though all of the components we manufacture currently are made to order, we do have a limited stock of some special components which might just happen to fit your requirements.

In any event, although our engineering staff, laboratories, and fully equipped shop are busy with government contracts, we will be happy to talk with you about your needs on similar work.

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DESIGN • RESEARCH • PRODUCTION
Microwave Transmission Lines and Associated Components
16 West 61st St. • New York 23, N. Y. • Circle 6-4760
TRADE MARK

## SECTION MEETINGS

*(Continued from page 38A)*

### PRINCETON

Symposium on the Problems Facing Young Engineers, by Martin Astrow, Sperry Gyroscope Company, M. W. Carpenter, RCA Laboratories, R. B. Hartman, II, Remington Arms Company, and A. S. Westneat, Jr., Applied Science Corporation; February 15, 1951.

### SACRAMENTO

"New Techniques in Single Side Band Transmission," by O. G. Villard, Jr., Faculty, Stanford University; May 10, 1950.

"Airport Communications and Navigational Aids Facilities," by F. H. McDonald and W. C. Hill, Sacramento Municipal Airport; June 27, 1950.

### SAINT LOUIS

"The Atomic Clock," by H. Lyons, National Bureau of Standards; November 30, 1950.

"Magnetic Amplifiers," by W. J. Dornhoefer Vickers Electric Division; December 21, 1950.

"Radar Landing Systems," by Paul Kreager, Bendix Radio Corporation; January 25, 1951.

"Television Facilities of the Bell System," by F. R. Small, Southwestern Bell Telephone Company; February 22, 1951.

"Fundamental Principles of Vibrator Power Supply Design," by I. M. Slater, P. R. Mallory & Company; March 22, 1951.

### SALT LAKE

"Allocation Engineering," by J. W. Wright, Columbia Broadcasting Company; March 19, 1951.

"Development of Simplified Single-Sideband Transmitting and Receiving Equipment for Voice and Teletype Service," by O. G. Villard, Jr., Faculty, Stanford University; April 10, 1951.

### SAN ANTONIO

"The Phenomenology of the Atomic Bomb," by J. E. Pickering, USAF School of Aviation Medicine, and "Electronics in Radiation Measurements," by W. J. Hamm, Faculty, St. Marys University; March 12, 1951.

### SAN FRANCISCO

"Television Picture Quality," by S. W. Athey, Faculty, Stanford University; December 13, 1950.

"A Half Million Volt RF Cascade Rectifier," by J. R. Woodyard, Faculty, University of California; January 4, 1951.

"Atomic Bomb Effects," by A. M. Dunn, United States Army; January 11, 1951.

"Electronic Research in European Laboratories," by Samuel Silver, University of California, L. A. Manning, Stanford University, and R. A. Helliwell, Stanford University; January 17, 1951.

### SCHENECTADY

"The Future of Color Television," by D. G. Fink, Editor, *Electronics;* April 11, 1951.

### SYRACUSE

"Phase Distortion in Audio Systems," by L. A. de Rosa, Federal Telecommunication Laboratories; April 5, 1951.

### TORONTO

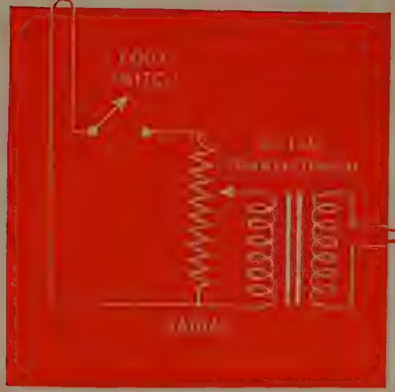
Student Night Speakers, A. E. Ticknor, I. R. Dutton, and J. C. Rennie, Students, University of Toronto; February 19, 1951.

"Multiplex Microwave Communication Equipment for an 1,800 Mile Pipe Line," by L. W. Goostree, General Electric Company; March 12, 1951.

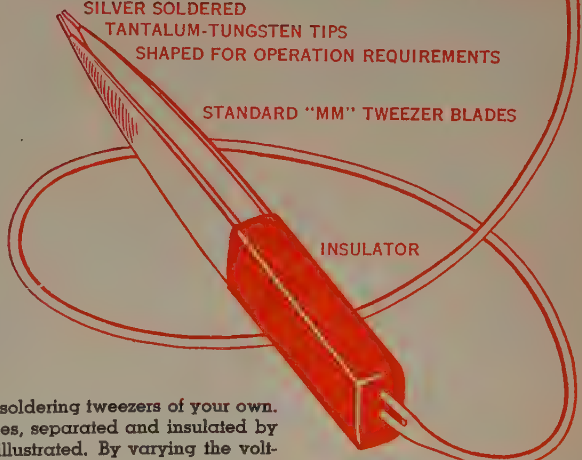
"Some Recent Developments in Audio Equipment," by J. K. Hilliard, Altec Lansing Corporation; March 28, 1951.

*(Continued on page 42A)*

# marion methods *electrify* TWEEZER SOLDERING


FOOT SWITCH

**Electrified by Marion, these tweezers now do *double duty. They position the work and solder it at the same time. Tripled* production and finer fluxless joints result.**


FLEXIBLE BRAIDED CONDUCTORS.
APPROX. EQUIV. #10 B. & S.
SILVER SOLDERED
TANTALUM-TUNGSTEN TIPS
SHAPED FOR OPERATION REQUIREMENTS
STANDARD "MM" TWEEZER BLADES
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**MARION USES THIS METHOD** of Tweezer Soldering for precise hairspring soldering, positioning small metal parts in thermoplastic cements, and in other fine soldering operations. Resistance heating in the work itself and conducted heat from the tips combine to give greatly increased soldering rates on pretinned components. Because the tweezers develop little heat themselves, adjacent component damage is avoided. These tweezers permit an inexperienced worker to rapidly acquire a high degree of skill. Both positioning and soldering are combined with the handling ease of a fountain pen.

**NOT FOR SALE**—You can make electric soldering tweezers of your own. Those shown here have standard "MM" blades, separated and insulated by bakelite. They are assembled and wired as illustrated. By varying the voltage, the tip cross-section, and the contact area, you can adapt this tool to your specific application.

**OTHER MARION METHODS** — Current demands on industry by the mobilization program accentuate the importance of production methods. Electric tweezer soldering is only one of a number of methods which Marion proposes to present in the hope that some of them can help you as much as they have helped us. Marion Electrical Instrument Company, 407 Canal Street, Manchester, N. H., U. S. A.

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MANUFACTURERS OF MARION RUGGEDIZED PANEL INSTRUMENTS

**marion meters**

Test Equipment FOR RADAR and PULSE APPLICATIONS

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Two additional R. F. Heads are now available for the Model 708 Spectrum Analyzer shown below:

- S Band 2500 to 3400 megacycles.
- $X_B$ Band 6200 to 7100 megacycles.

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MODEL 708 SPECTRUM ANALYZER

Frequency range—8500 mc to 9600 mc.
Receiver—Double conversion superheterodyne.
IF bandwidth—approximately 10 kc.
Sweep frequency—10 cps to 25 cps.
Minimum frequency dispersion—1 mc/inch.
Maximum frequency dispersion—10 mc/inch.
Signal input attenuator—100 db linear.
Power—115V or 230V, 50 cps to 800 cps.

Write for complete technical data

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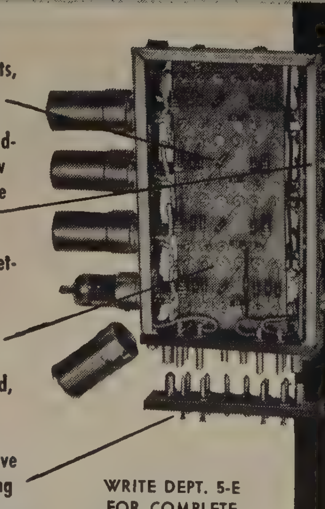
Simple • Reliable • Economical
Potter decimal counter



Highest quality pretested components, conservative ratings

Four large, easy reading, bulls-eye glow lamps – replaceable socket type

All components turret-lug mounted and accessible . . . all wiring color coded

Special silver plated, self-aligning contact and rigid connectors for positive mechanical mounting

DIRECT DECIMAL READ-OUT — FOUR NEON GLOW LAMPS DESIGNATED 1-2-4-8 PROVIDE DIRECT INDICATION (0-9) AND INSTANTANEOUS LOCATION OF ANY DEFECTIVE TUBE.

STABLE OPERATION — WIDE VOLTAGE RANGE.

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A NEW LOW UNIT PRICE OF \$45.00 IS ANNOUNCED AS A RESULT OF WIDE ACCEPTANCE AND QUANTITY PRODUCTION

POTTER INSTRUMENT CO.
INCORPORATED
115 CUTTER MILL RD., GREAT NECK, N.Y.

## SECTION MEETINGS

(Continued from page 40A)

### TWIN CITIES

"Predicting Radio Propagation," by F. J. Maclary, Aeronautical Radio, Incorporated; February 22, 1951.

"At Your Service: Electric Brains and Wire Nerves," by R. E. Nelson, Bell Telephone Company; March 6, 1951.

### WASHINGTON

"The Aerobee and High Altitude Radio Reporting," by G. H. Melton, Johns Hopkins University; March 12, 1951.

## *SUBSECTIONS*

### BINGHAMTON

"Micro Wave Relay Systems," by H. U. Lotz, American Telephone and Telegraph Company; March 13, 1951.

### HAMILTON

"Some Aspects of Phonograph Reproduction," by A. B. Oxley, RCA Victor Company; April 9, 1951.

### LONG ISLAND

"Transformer Concepts for Simplifying Networks," by H. A. Wheeler, Wheeler Laboratories; March 13, 1951.

### NORTHERN NEW JERSEY

Tour of Station WOR-TV; March 14, 1951.

### PALO ALTO

Election of Officers and Talk: "Digital Frequency Measurements up to 10 Megacycles per Second," by Allen Bagley, Hewlett-Packard Company; February 6, 1951.

## STUDENT BRANCH MEETINGS

### CLARKSON COLLEGE OF TECHNOLOGY, IRE BRANCH

Election of officers and two films; "Basic Principles of Frequency Modulation" and "The Effects of the Ionosphere on Radio Wave Propagation"; April 19, 1951.

### UNIVERSITY OF COLORADO, IRE-AIEE BRANCH

Nomination of officers, discussion of two prospective student tours, and AIEE Convention in Lincoln, Nebraska, and film; "The World Series of 1949," April 4, 1951.

### UNIVERSITY OF CONNECTICUT, IRE-AIEE BRANCH

"Magnetrons," by Edgar Evenhart, Faculty, University of Connecticut and nomination of officers; April 19, 1951.

### UNIVERSITY OF DENVER, IRE-AIEE BRANCH

"Selenium Rectifiers," by W. O. Pratt, "Microwave Transmission," by A. N. Clark, "Electronics in Industry," by Anwar Chitayat, and "Germanium Diodes," by L. T. Kelly, Students, University of Denver; March 10, 1951.

### DREXEL INSTITUTE OF TECHNOLOGY, IRE-AIEE BRANCH

General meeting and four films; "Speeding Speech," "A Million Times a Day," "Rehearsal," and "Kingdom of Plastics"; March 8, 1951.

(Continued on page 44A)

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picture
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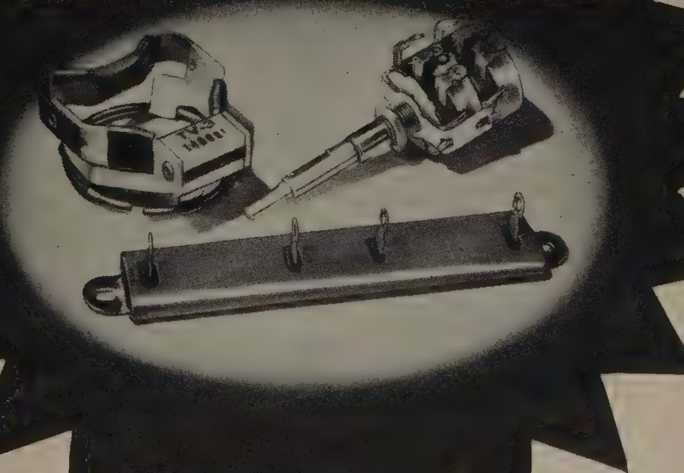
**Television**
is . . .

CLAROSTAT

FOR CONTROLS,
RESISTORS, BEAM-BENDERS



As "Big Business," television came suddenly. Just as suddenly, Clarostat was ready. Three decades of pioneering and specialization, backed by a plant second to none, assured TV designers and manufacturers of an outstanding selection of resistors, controls and resistance devices.

And when ion spot blemishes became a major problem, again Clarostat was ready with simpler and cheaper beam-benders.

Thus Clarostat products are already represented in over 5,000,000 sets and in countless radios in daily use. All because, for quality, uniformity, dependability, economy, it's CLAROSTAT.

Controls and Resistors

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IN CANADA: CANADIAN MARCONI CO., LTD. MONTREAL, P. Q., AND BRANCHES

CLAROSTAT

What LORD RESEARCH in Vibration-Control means to YOU

• This modern testing machine measures the deflection of LORD Vibration-Control Mountings under load and autographically records the deflection curve. Although deflection testing is but one phase of an extensive program, it serves to illustrate the precision and excellent facilities which are found in all branches of LORD research.

Continuous investigation of metals, elastic materials, bonding methods, and mounting performance have resulted in improvements of value to every user of vibration-control mountings and custom made rubber-bonded-to-metal parts. LORD has developed over 500 natural and synthetic rubber stocks. From these is selected the one with the *exact* characteristics to deliver maximum performance and longest life for each application. A choice of metals means adequate strength . . . minimum weight . . . maximum corrosion resistance. New bonding methods improve quality and lower cost.

Research has made it possible for LORD to produce more accurate . . . uniform . . . dependable . . . economical vibration-control mountings and bonded-rubber parts. LORD Field Engineering Representatives are ready to assist with proper selection and application. Write for your copy of the Lord Natural Frequency Chart and of the Vibration Isolation Chart. Designers and engineers will find them of definite value.

**LORD MANUFACTURING COMPANY • ERIE, PA.**
Canadian Representative: Railway & Power Engineering Corp. Ltd.

LORD BONDED RUBBER

Vibration-Control Mountings
. . . Bonded-Rubber Parts

## STUDENT BRANCH MEETINGS

*(Continued from page 42A)*

FENN COLLEGE, IRE BRANCH

"Fundamentals of Radar," by Leo M. Beckerman, Student, Fenn College; April 20, 1951.

UNIVERSITY OF FLORIDA, IRE-AIEE BRANCH

"Hurricane Tracking by Radar," by M. H. Latour, Faculty, University of Florida; April 3, 1951.

GEORGE WASHINGTON UNIVERSITY, IRE BRANCH

"Radio Astronomy," by Grote Reber, Atmospheric Research Division, National Bureau of Standards, and Election of Officers; November 1, 1950.

"Color Television," by William L. Everitt, Faculty, University of Illinois; December 6, 1950.

"Frequency Standards at WWV," by George Resser, National Bureau of Standards; January 11, 1951.

Films; "Microwave Oscillators," and "Fundamentals of Frequency Modulation"; March 7, 1951.

Films; "The Cathode Ray Tube," "Radio Transmitters and Circuits," and "Ionosphere Effects on Radio Propagation"; April 4, 1951.

GEORGIA INSTITUTE OF TECHNOLOGY, IRE BRANCH

Film; "Electrical Proving Ground"; April 3, 1951.

Business Meeting; April 10, 1951.

UNIVERSITY OF KENTUCKY, IRE BRANCH

"Radio Astronomy," by M. Davis, Faculty, University of Kentucky; March 22, 1951.

"Servoanesthetizers," by E. J. Breiding, Student, University of Kentucky; April 5, 1951.

Film; "Watts in Glass"; April 12, 1951.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH

"Television," by J. A. Feron, Student, Lafayette College; April 10, 1951.

LEHIGH UNIVERSITY, IRE BRANCH

Field trip through Federal Telecommunications Laboratories, Nutley, New Jersey; March 16, 1951.

LOUISIANA STATE UNIVERSITY, IRE-AIEE BRANCH

"Student Convention," by W. J. Seeley, Faculty, Duke University and film, "Electrical Proving Ground"; March 27, 1951.

MARQUETTE UNIVERSITY, IRE-AIEE BRANCH

Election of officers; April 5, 1951.

"Boiler Design and Construction," by D. D. Smith, Babcock, and Wilcox; April 19, 1951.

UNIVERSITY OF MARYLAND, IRE-AIEE BRANCH

Election of officers; April 18, 1951.

UNIVERSITY OF MIAMI, IRE BRANCH

"Modern System and Equipment Developments for Adjacent or Split Channel Communications," by L. P. Morris, Motorola Corporation; March 30, 1951.

Discussion of methods to increase membership of IRE Student Branch; April 9, 1951.

UNIVERSITY OF MICHIGAN, IRE-AIEE BRANCH

"Electric Telegage," by L. J. Vanderberg, King Seely Corporation; March 26, 1951.

MICHIGAN COLLEGE OF MINING AND TECHNOLOGY, IRE-AIEE BRANCH

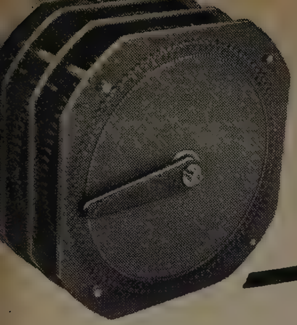
"Television," by Nels Johnson, Philco Radio Corporation; February 8, 1951.

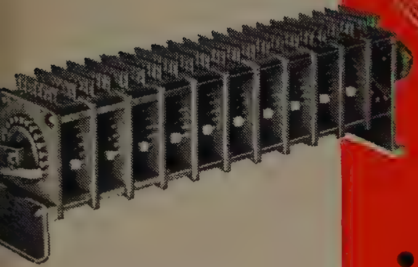
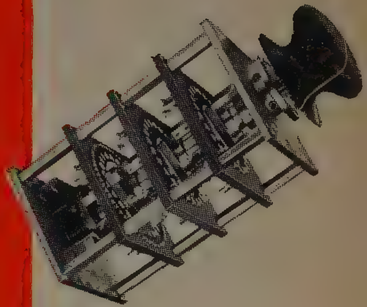
General meeting; February 20, 1951.

Discussion of forthcoming Engineering Show; March 6, 1951.

*(Continued on page 46A)*

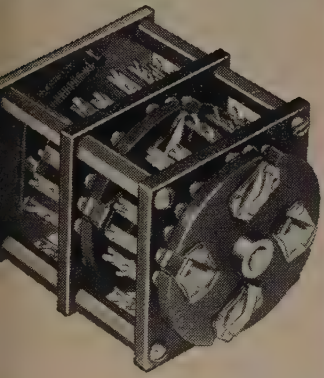
# Quick Switches



## are a

# DAVEN
## SPECIALTY



And the "specialty of the house" is double-barreled . . . first, choose from hundreds of standard units to satisfy your needs—for quick switch delivery . . . second, Daven can effect quick 'switches' or changes from standard units to special switches, by using components at hand. Standard parts can be adapted for *your* switch. That too makes for speed, dependability, economy. Write for more detailed data.

## Here's Why Daven Switches Excel

- Low and uniform contact resistance.
- Minimum thermal noise.
- High resistance to leakage.
- Trouble-free operation and long life.
- Roller-type positive detent action.
- Depth of unit not increased by addi-

Standard Daven Switches may be the answer to many of your problems. Therefore, check this list below for many of the popular types that are readily available.

| Type | Operation | Maximum No. of Positions (per pole) | Maximum Poles per Deck | Deck |
|---|---|---|---|---|
| G1A | Make before break | 24 | 1 | 1 3/8" |
| C1A | Make before break | 31 | 1 | 1 3/4" |
| C2B | Break before make | 15 | 1 | 1 3/4" |
| D1A | Make before break | 47 | 1 | 2 1/4" |
| D7A | Make before break | 14 | 4 | 2 1/4" |
| D8B | Break before make | 7 | 4 | 2 1/4" |
| D9A | Make before break | 9 | 5 | 2 1/4" |
| E3A | Make before break | 47 | 2 | 2 3/4" |
| E8B | Make before break | 12 | 4 | 2 3/4" |
| E11A | Make before break | 15 | 6 | 2 3/4" |
| F1A | Make before break | 60 | 1 | 3" |

SWITCHES
BY
DAVEN
THE DAVEN CO.

## It's Free!

Your copy of Daven's complete, new bulletin on switches.

THE DAVEN CO.
NTRAL AVENUE • NEWARK 4, NEW JERSEY

FROM 3 INCHES TO 10 FEET

"S" Band Radar Antennas for the Armed Forces

Nothing could better demonstrate the versatility of the WORKSHOP staff and facilities than these two radar antennas designed for the Army Signal Corps. The 3-inch gold-plated "beacon," flush-mounted on the belly of a plane, is "fired" by the 10-foot parabolic antenna . . . they operate as a team. This extreme range in size is easily accommodated by the WORKSHOP's elaborate research and test equipment.

Military antennas of all types — rocket, guided missile, beacon and communications — are being developed regularly in the WORKSHOP laboratories under Government contracts. Our engineering staff is always available for consultation on any antenna problem without obligation. Write, or phone Needham 3-0005.

THE WORKSHOP ASSOCIATES
Incorporated
*Specialists in High-Frequency Antennas*
135 Crescent Road, Needham Heights 94, Mass.

# STUDENT BRANCH MEETINGS

*(Continued from page 44A)*

UNIVERSITY OF MINNESOTA, IRE-AIEE BRANCH

"Reduction of Noise in Receivers," by A. van der Ziel, Faculty, University of Minnesota; April 4, 1951.

UNIVERSITY OF MISSOURI, IRE-AIEE BRANCH

"A Rapid Recovery Limiting Amplifier Featuring Low Distortion," by J. B. Harvey, Student, University of Missouri; March 27, 1951.

Election of officers; April 10, 1951.

MISSOURI SCHOOL OF MINES AND METALLURGY, IRE-AIEE BRANCH

"If I Were You," by J. Hoffman, James R. Kearney Corporation, and election of officers; March 1, 1951.

"House of Magic"; March 13, 1951.

Student Papers Contest; March 29, 1951.

COLLEGE OF THE CITY OF NEW YORK, IRE BRANCH

"Microwave Relay Transmission," by Mr. Polasius, Bell Telephone Laboratories; March 22, 1951.

"Panoramic Adapters," by Mr. Schlessel, Sales Engineer of Panoramic; April 12, 1951.

NEW YORK UNIVERSITY (DAY DIV.), IRE BRANCH

"Acoustical Analogies," by C. F. Rehberg, Faculty, New York University; March 29, 1951.

Film; "Atomic Physics"; April 2, 1951.

"Feedback Amplifiers," by Dr. Mulligan, Faculty, New York University; April 5, 1951.

NEW YORK UNIVERSITY (EVENING DIV.), IRE-AIEE BRANCH

"Electrical Interconnections Between Power Plants On The Eastern Seaboard," by J. Gladdis, Ebasco Services, Incorporated; March 28, 1951.

OHIO STATE UNIVERSITY, IRE-AIEE BRANCH

"Betatron and DeGraf Generator," by C. J. Carter, Faculty, Ohio State University; April 5, 1951.

UNIVERSITY OF PENNSYLVANIA, IRE-AIEE BRANCH

Business meeting; April 16, 1951.

PRATT INSTITUTE, IRE BRANCH

"Factors in the Design of Vacuum Tubes," by Mr. Gaenzle, Sylvania Electric Company; March 28, 1951.

PRINCETON UNIVERSITY, IRE-AIEE BRANCH

"Electrical Engineering Applications in Geophysics," by J. C. Maxwell, Faculty, Princeton University; March 15, 1951.

PURDUE UNIVERSITY, IRE BRANCH

"Analog Computers," by D. C. Ross, Faculty, Purdue University; April 19, 1951.

UNIVERSITY OF SOUTH CAROLINA, IRE BRANCH

Election of officers; January 10, 1951.

"Inductively Coupled Circuits," by J. P. Doff, "Effects of Harmonics on Polyphase Transformations," by W. D. Rohe, "Electrostatic Storage" or "Memory Tubes," by W. T. Price, and "Civil Aeronautics System of Omnirange Navigation," by G. M. Crook, Students, University of South Carolina; January 15, 1951.

General meeting and film on development of the Atomic Bomb; February 9, 1951.

"Television Antennas," by H. E. Eidson, Radio Station WIS; February 21, 1951.

UNIVERSITY OF TEXAS, IRE-AIEE BRANCH

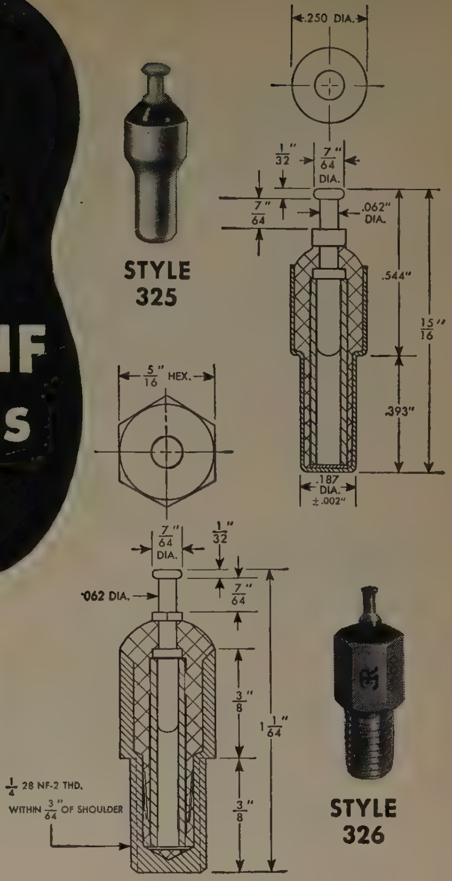
"A Survey of Structural Solutions by Electrical Analogies," by R. M. Westkaemper, Student, University of Texas; April 2, 1951.

*(Continued on page 48A)*

## ERIE STYLES 325 and 326 STAND-OFF CERAMICONS®

# The Most Efficient Stand-Off Units for VHF and UHF Applications

## HERMETICALLY SEALED . . . RUGGED



BETTER by-passing is one solution to control of regeneration in high frequency receivers. The Erie Styles 325 and 326 Stand-off Ceramicons offer outstanding features for extremely efficient by-passing. Style 325 is designed for push-on clip mounting to chassis; Style 326 is for threaded mounting installation.

This basic design provides, in a hermetically sealed case, a by-pass-to-ground through the shortest possible path. Full advantage is taken of the concentric electrode configuration in maintaining this short path by making connection to the outer electrode at the plane of the chassis. The result is an extremely low and uniform series inductance. In assembly operations both location and length of leads are accurately fixed, resulting in better mechanical uniformity. This feature has been demonstrated to be of particular importance in good VHF and UHF design.

### OUTSTANDING FEATURES

① Tubular ceramic capacitor is completely sealed.

② Very low and uniform inductance path to ground.

③ Electrical shielding is provided by means of the grounded metal case.

④ Post terminal provides a sturdy tie point for several connections, and is made essentially to match tube socket terminal height in the interest of maintaining uniform short leads.

### SPECIFICATIONS

Standard available capacitance values in MMF: 10, 33, 47, 68, 82, 100, 680, 1,000, 1,500. Voltage Rating, 500 VDC.

*Electronics Division*

**ERIE RESISTOR CORP., ERIE, PA.**

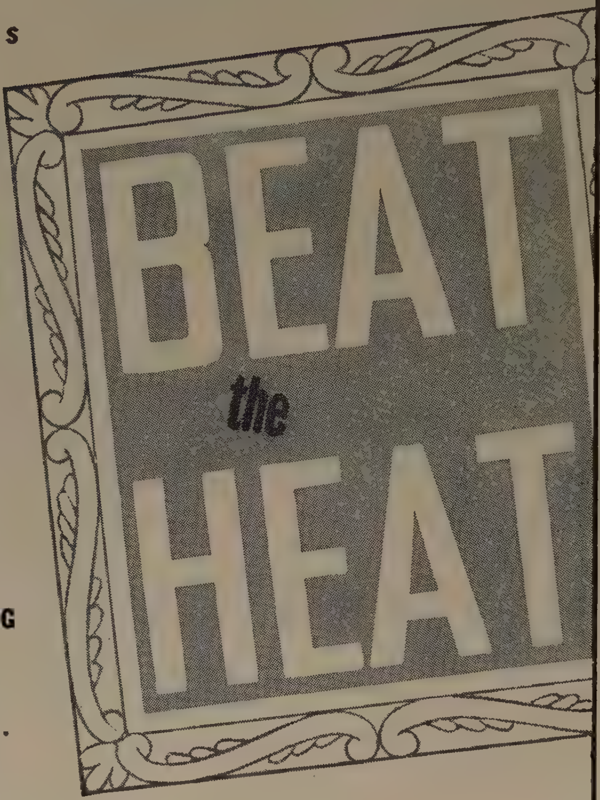
LONDON, ENGLAND . . . TORONTO, CANADA

APPROVED BY UNDERWRITERS LABORATORIES AT 90°

"NOFLAME-COR"

the TELEVISION hookup wire

fits perfectly
into this
picture



Meets government specification AN-J-C-48a

FLAME RESISTANT
HEAT RESISTANT
HIGH INSULATION RESISTANCE
EASY STRIPPING
HIGH DIELECTRIC
FACILITATES POSITIVE SOLDERING

*Also*
is not affected by the heat of impregnation . . . making it an ideal wire for use in connection with coil and transformer leads

CORNISH
"made by engineers for engineers"

No "Nicking" problem in using this proven wire. Not being an extruded plastic, its diameter uniformity can be absolutely guaranteed. This eliminates nicking of conductors and constant resetting of blades in the cutting process. Available in all sizes, solid and stranded, in over 200 color combinations . . . "NOFLAME-COR" assures maximum output and minimum rejects.

*No "blobbing" of insulations under soldering heat, because "NOFLAME-COR" is NOT an extruded plastic. Production executives specify it as the most efficient heat-resistant wire yet developed. Save time, money and assembling headaches. Investigate!*

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DALLAS DENVER LOS ANGELES SAN FRANCISCO SEATTLE

MANUFACTURERS OF QUALITY WIRES AND CABLES FOR THE ELECTRICAL AND ELECTRONIC INDUSTRIES

## Student Branch Meetings

*(Continued from page 46A)*

"Determination of Resistance to Ground of Grounding Grids," by A. J. McCrocklin and C. W. Wendlandt, Faculty, University of Texas; April 16, 1951.

UNIVERSITY OF VIRGINIA, IRE BRANCH

"Voltage Regulators," by Ray Johnson, and "Magnetostriction Dentist Drill," by Jim Holyman, Students, University of Virginia; March 13, 1951.

"High Voltage Generators," by Ken Kyle, and "Analog Computers," by F. Cuddihy, Students, University of Virginia; March 27, 1951.

VIRGINIA POLYTECHNIC INSTITUTE,
IRE-AIEE BRANCH

Film; "The World's Greatest Workshop"; April 17, 1951.

UNIVERSITY OF WASHINGTON, IRE-AIEE BRANCH

Talk by R. A. Turnbull; January 16, 1951.

"Industrial Transformer Testing," by L. B. Robinson, General Electric Company; January 30, 1951.

"Employment Opportunities in the Field of Servomechanisms," by T. M. Stout, Faculty, University of Washington"; February 6, 1951.

"Student Activities," by J. A. McDonald, Vice-President of the Ninth District, American Institute of Electrical Engineers; February 19, 1951.

"Look At Your Opportunities Objectively," by H. Muller, Westinghouse Manufacturing Company; February 28, 1951.

"Bonneville Power Administration—Its Organization and Major Engineering Problems," by O. A. Demuth, Bonneville Power Administration; March 6, 1951.

Election of officers; April 3, 1951.

"The Problem of Electric Power in the Modern World," by J. T. Wilson, Allis-Chalmers Manufacturing Company; April 18, 1951.

UNIVERSITY OF WISCONSIN, IRE BRANCH

"AIEE District Meeting," by J. D. Baird, Faculty, University of Wisconsin; February 15, 1951.

WORCESTER POLYTECHNIC INSTITUTE,
IRE-AIEE BRANCH

"Adventures in Engineering," by Louis Leavitt, Worcester County Electric Company; March 22, 1951.

UNIVERSITY OF WYOMING, IRE-AIEE BRANCH

General meeting; March 13, 1951.
General meeting; April 3, 1951.

YALE UNIVERSITY, IRE-AIEE BRANCH

"The Design of the Viking Ballistic Rocket," by W. B. Bergen, Glenn L. Martin Company; March 1, 1951.

20,420
Radio Engineers
"look it up"
in the
IRE DIRECTORY

# Go on the Air to Stay

## with JOHNSON

ANTENNA PHASING EQUIPMENT

ANTENNA AND TRANSMITTER

COMPONENTS

**Typical Johnson Antenna Phasing Equipment**

Hundreds of Johnson phasing installations are on the air — to stay — for you can depend on JOHNSON for the best! Designed especially for your station, and incorporating the recommendations of your consulting engineers, JOHNSON phasing equipment offers a host of advantages. It has optimum circuit design, heavier components, automatic switching from directional to non-directional operation, and others.

### OTHER JOHNSON BROADCAST EQUIPMENT

- Adjustable Phase Sampling Loops
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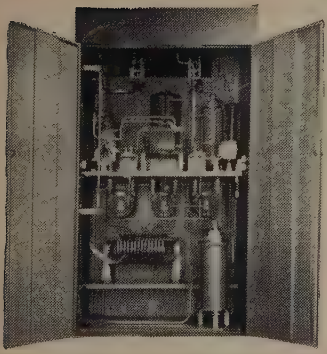
- Variable Capacitors
- Standing Wave Indicators
- R F Contactors
- Tower Lighting Filters
- Transmission Line Supports
- Pressurized Capacitors
- Neutralizing Capacitors
- Fixed Inductors
- Variable Inductors
- Feed-Thru Bowl Assemblies
- Make Before Break Switches

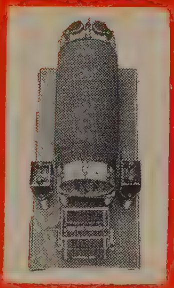
*Write for specific information directly or through your consulting engineer.*



**Interior view of Phasing Equipment**

**R F Contactors**

**Tower Lighting Filter**



**Johnson Coupling Unit**

**Johnson Adjustable Shielded Sampling Loop**

**JOHNSON** *a famous name in radio*

**E. F. JOHNSON CO.** **WASECA, MINNESOTA**

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DEVELOPMENT
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*Career Positions*
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*in*
Radar Pulse, Timing and
Indicator Circuit Design
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Also
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Experience in servo-mechanism, special weapons, fire control, and guided missile design.

Recent E.E. graduates and those with at least one year electronics research and development work will also be considered.

Salary commensurate with ability. Housing reasonable and plentiful.

Submit resume outlining qualifications in detail. Information will be kept strictly confidential. Personal interviews will be arranged

THE GLENN L. MARTIN COMPANY
Employment Department
Baltimore 3, Maryland

**POSITIONS AVAILABLE**
GRADUATE
**ELECTRONIC ENGINEERS**
EXPERIENCE IN DESIGN
AND DEVELOPMENT PREFERRED
**FOR PROJECTS**
**IN THE FIELDS OF**
**"TELEMETERING"**
**AND**
**"RADIO CONTROL"**
PROJECTS WITH EXCELLENT
PEACETIME APPLICATION

Give experience, education, age, salary expected. Please be complete and specific. All inquiries will be considered promptly and kept confidential.

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PHILADELPHIA 4, PA.

# POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . .

**The Institute reserves the right to refuse any announcement without giving a reason for the refusal.**

PROCEEDINGS of the I.R.E.
I East 79th St., New York 21, N.Y.

### ENGINEERS

Precision instrument plant has positions available for highest caliber personnel in the field of airborne automatic electro-mechanical control equipment. Mechanical design, electronic, servo and Junior Engineers. Write or reply: AC Spark Plug Division, General Motors Corp., 1925 E. Kenilworth Place, Milwaukee 2, Wisconsin.

### ELECTRICAL ENGINEER—PHYSICISTS

Men to train in oil exploration for the operation of seismograph instruments and the computing and interpreting of seismic data. Beginning salary $250 to $325 depending upon background. Excellent opportunity for advancement determined by integrity and ability. Nature of work requires several changes of address each year; work indoors and out; general location in oil producing states. To apply, write giving scholastic and employment background; age, nationality, marital status, and include recent snapshot to National Geophysical Co., Inc., 8800 Lemmon Ave., Dallas, Texas.

### ELECTRICAL ENGINEERS

Electrical engineers to design sound equipment, audio amplifiers and electric carillons. Requirements: B.S. degree, 1 to 5 years' experience in audio, electronic and acoustical systems. Location: upstate New York. Box 653.

### ELECTRICAL ENGINEER

New Florida research company requires engineer preferably with nuclear instrumentation experience in tubes or circuits. Part or full time with age no objective. Write qualifications in full to Box 951, West Palm Beach, Florida.

### ENGINEERS—TECHNICIANS

Large midwest manufacturer has positions open for electrical and electronics engineers and technicians. Applicants are required to submit data concerning their education and experience together with photograph. Box 656.

### ENGINEERS—PHYSICISTS

Several Junior and Senior engineers are required immediately for the direction of important research and design work by an organization specializing in all types of advanced armament development. This is an opportunity to participate in the growth of an exceptionally active group in a highly technical field. Positions open include: electronics engineers, electromechanical engineers, engineering physicists. Write, phone or wire to R. W. Sanford, Ch. Eng. Aircraft Armaments, Inc. 4415 Reistertown Rd., Baltimore 15, Md.

### ELECTRICAL ENGINEERS-MECHANICAL ENGINEERS

Bendix Aviation Corporation, Kansas City Division has openings for electrical engineers and mechanical engineers with at least 3 years' design and development experience in electrical communications or navigation equipment. Particularly interested in design and development of video circuits, transmission lines and antennas. Airmail complete details of qualifications to Bendix Aviation Corporation, Kansas City Division, Kansas City, Missouri.

*(Continued on page 52A)*

**WANTED**
Radio and Electronic
Engineer

with background in RF coils to work for coil manufacturer in Southern California. Man must be of highest caliber, and able to assume responsibilities and take charge of Engineering Department.

Salary commensurate with ability.

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The Institute of Radio
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**ENGINEERS**
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**Microwave . . .**
to develop test sets and methods on long-distance radio relay equipment. Experienced in performance of amplifiers, modulators, filters and cavities at microwave frequencies.

**Filters and Networks . . .**
to develop manufacturing facilities and methods of manufacture.

**Test . . .**
to develop test facilities for the manufacture of communication and television transmission. Basic knowledge of transmission theory required.

Write Dept. P giving details
of qualifications, experience
and salary expected or
APPLY IN PERSON
EMPLOYMENT DEPARTMENT
Monday through Saturday,
8:30 A.M. to 3:30 P.M.

Applicants must furnish proof of citizenship and social security number. Veterans must present discharge papers.

**WESTERN ELECTRIC CO.**
100 Central Ave., Kearny, N.J.
SUPPLY UNIT OF THE BELL SYSTEM

# Physicists
# Engineers

The Boeing Airplane Company has exceptional openings at Seattle for outstanding physicists and engineers. Are you one of them?

The work is in Boeing's expanded Physical Research Unit—on some of the nation's most vital projects: research and development on

**Weapon systems**
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These positions offer challenging work and a stimulating professional environment. You'll live in the Pacific Northwest with its equable climate and unparalleled recreational possibilities.

Opportunity for advancement is essentially unlimited in the rapidly expanding fields of guided missiles, airplane control and industrial machine and process control. Salaries are based on semi-annual performance review.

To qualify, you must already have demonstrated outstanding experimental or analytical abilities. Or you must have recently received the M.S. or Ph.D. degree in Electrical Engineering, Physics, or Applied Mathematics.

If you qualify, write today for application. . . .

JOHN C. SANDERS, *Staff Engineer–Personnel*
*Boeing Airplane Company*
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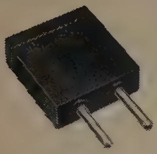
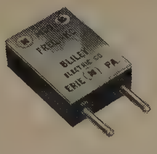
## SPECIAL OPENINGS FOR ENGINEERS

*Boeing's Engineering Division needs experienced and junior engineers in the following and allied engineering fields: mechanical, aeronautical, electrical, electronic, civil, acoustical, and structural. Write today for application forms to address indicated above, right.*

**BOEING**

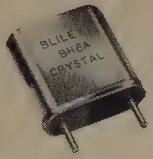
Q
M
S?


TYPE SR5A
FREQ 2.0-15.0 MC



TYPE MC9
FREQ 1.0-10.0 MC

TYPE AR23W
FREQ
0.080-0.19999 MC



TYPE BH6A
FREQ 0.8-75.0 MC

TYPE BH7A
FREQ 15.0-50.0 MC

TYPE TCO-1
TEMPERATURE
CONTROL OVEN

Questions
Concerning Military Specifications?

Bliley is well acquainted with "MIL" crystal requirements. Solid production experience is an important factor when you need "MIL" *quality* as well as *dependable* delivery.

Bulletin 42, describing "MIL" crystals, will be sent to *design* engineers on request.

Bliley
CRYSTALS

**BLILEY ELECTRIC COMPANY**
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**ERIE, PENNSYLVANIA**

**RESEARCH OPPORTUNITIES IN THE LOS ANGELES AREA**

**Unusual Opportunity for Senior men with degrees and at least five years of outstanding proven accomplishment to achieve further growth by working with some of the nation's outstanding scientists on commercial and military projects in large modern electronics laboratories.**

| | |
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| ELECTRONIC ENGINEERS | SERVOMECHANISMS ENGINEERS |
| PHYSICISTS—CIRCUITRY | ELECTRO MECHANICAL ENGINEERS |
| PHYSICISTS—ANALYSIS | MECHANICAL DESIGNERS |
| PHYSICISTS—OPTICS | |
| PHYSICISTS—ELECTRON TUBES | |

**Long term program of research and development in the fields of Radar, Guided Missiles, Computers, Electron Tubes, and related equipment.**

***Please do not answer unless you meet the above requirements.***

**RESEARCH AND DEVELOPMENT LABORATORIES**
**Hughes Aircraft Company** **CULVER CITY, CALIFORNIA**

**OPENINGS FOR**
**ELECTRICAL ENGINEERS**
**MECHANICAL ENGINEERS**

**Requires minimum of 3 years' design and development experience in electrical communications or navigation equipment. Particularly interested in design and development of video circuits, transmission lines and antennas.**

***Airmail Complete Details of Qualifications***

BENDIX AVIATION
CORPORATION
KANSAS CITY DIVISION
KANSAS CITY, MISSOURI

## POSITIONS OPEN

*(Continued from page 50A)*

### ELECTRICAL ENGINEER-ELECTRONICS

At least 5 years' practical experience in instrumentation. Ph.D. preferred. Must have experience in servomechanisms, AC and DC amplifiers, audio and radio frequency equipment, measuring instruments and components. Salary open. Box 657.

### RADAR ENGINEERS

Electrical and mechanical engineers, preferably with experience in design, development or manufacturing engineering of airborne radar equipment. Circuit, systems, scanner design or waveguide phases of experience desired. Long term backlog of projects. Location southern California. Send Airmail detailed résumé and salary requirements to F. G. Suffield, Mgr. of Engineering, The Houston Corp. 11801 W. Olympic Blvd., West Los Angeles 64, Calif.

### FEMALE ELECTRICAL ENGINEER

Electronics experience preferred, correspondence and some typing. Excellent opportunity. Location: Brooklyn, N.Y. Box 659.

### ELECTRONICS ENGINEERS

Electronics Office, New York Naval Shipyard, Brooklyn, N.Y. has openings for men with college degrees and 1 or more years of experience in the design, test, calibration, or field evaluation of radar, communications, or sonar equipments. Starting salaries $3825 or 4600 according to experience. Address replies, containing a brief résumé of education and experience to Electronics Officer (Code 125) N. Y. Naval Shipyard, Brooklyn 1, N.Y. or call Main 5-4500. Ext. 520.

### DIRECTOR

Director wanted. Salary $100 per week. Must be high school graduate or equivalent. Must also qualify for any 2 of the following 3 items: (1) Journeyman for at least 2 years in radio and television. (2) 2 years teaching experience. (3) 1 year directing or administering a school. Write Tru-Way Radio & Television School, 231 Arch St., Nanticoke, Pennsylvania.

### ENGINEERS, RADIO AND TELEVISION

Several permanent positions open for engineers having various degrees of experience in radio and television receiver design. Salaries open depending upon experience and ability to applicant. Location: Newark, New Jersey. Send résumé to Box 660.

### ENGINEERS

Experience in following fields desirable: I.F. and R.F. coils, mica and ceramic capacitors, powdered metals, ceramics, television tuners and transformers, radiosondes, electronic equipment, and Government equipment. Positions are permanent and offer excellent opportunities. Salary open. Send résumé to Automatic Manufacturing Corp. 65 Gouverneur St., Newark, New Jersey.

### FACSIMILE ENGINEER

Electronic section of leading News Syndicate offers permanent position to electronic engineer for research and development work in facsimile. Location New York City. Reply Box 661.

### ENGINEERS—PHYSICISTS

Engineers, physicists: at least 5 years' experience on circuit design and development of digital and analog computers, servomechanims, oscilloscopes, radar, amplifiers and power supplies. Must be familiar with government specifications. Salary commensurate with experience. Location New York City. Box 662.

*(Continued on page 54A)*

## PHYSICISTS AND SENIOR RESEARCH ENGINEERS

### POSITIONS NOW OPEN

Senior Engineers and Physicists having outstanding academic background and experience in the fields of:

- Microwave Techniques
- Moving Target Indication
- Servomechanisms
- Applied Physics
- Gyroscopic Equipment
- Optical Equipment
- Computers
- Pulse Techniques
- Radar
- Fire Control
- Circuit Analysis
- Autopilot Design
- Applied Mathematics
- Electronic Subminiaturization
- Instrument Design
- Automatic Production Equipment
- Test Equipment
- Electronic Design
- Flight Test Instrumentation

are offered excellent working conditions and opportunities for advancement in our Aerophysics Laboratory. Salaries are commensurate with ability, experience and background. Send information as to age, education, experience and work preference to:

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Aerophysics Laboratory
Box No. N-4, 12214 South Lakewood Blvd.
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## ENGINEERS PHYSICISTS

A major expansion of THE JACOBS INSTRUMENT COMPANY has made available a number of positions for men qualified to take responsibility as Section Heads. A minimum of five years of experience, plus an outstanding record of achievement, is required. Experience should be in high frequency circuit design, systems studies, aircraft instruments, aerodynamics, servomechanisms, gyroscopes, optical instruments, missile guidance, fire control, digital or analog computers, or related fields.

These positions offer many unusual attractions including high salary, unusual opportunity in military and peacetime fields, security, employee benefits, and pleasant working conditions.

Qualified individuals interested in advancing themselves now should not overlook these opportunities.

**JACOBS**
**INSTRUMENT CO.**
Bethesda, Maryland

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## SCIENTISTS AND ENGINEERS

*for*

challenging research and advanced development in fields of

RADAR
GYROSCOPES
SERVOMECHANISMS
MECHANICAL SYSTEMS
ELECTRONICS CIRCUITS
APPLIED PHYSICS AND MATH
PRECISION MECHANICAL DEVICES
ELECTRICAL SYSTEM DESIGN
GENERAL ELECTRONICS
INSTRUMENTATION
MICROWAVES
COMPUTERS
AUTOPILOTS

*Scientific or engineering degree and extensive technical experience required.*

WRITE:
*Manager,* ENGINEERING PERSONNEL
**BELL AIRCRAFT CORPORATION**
P.O. Box 1, Buffalo 5, N.Y.

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## ENGINEERS . . .
## PHYSICISTS . . .
## CHEMISTS . . .
## METALLURGISTS . . .

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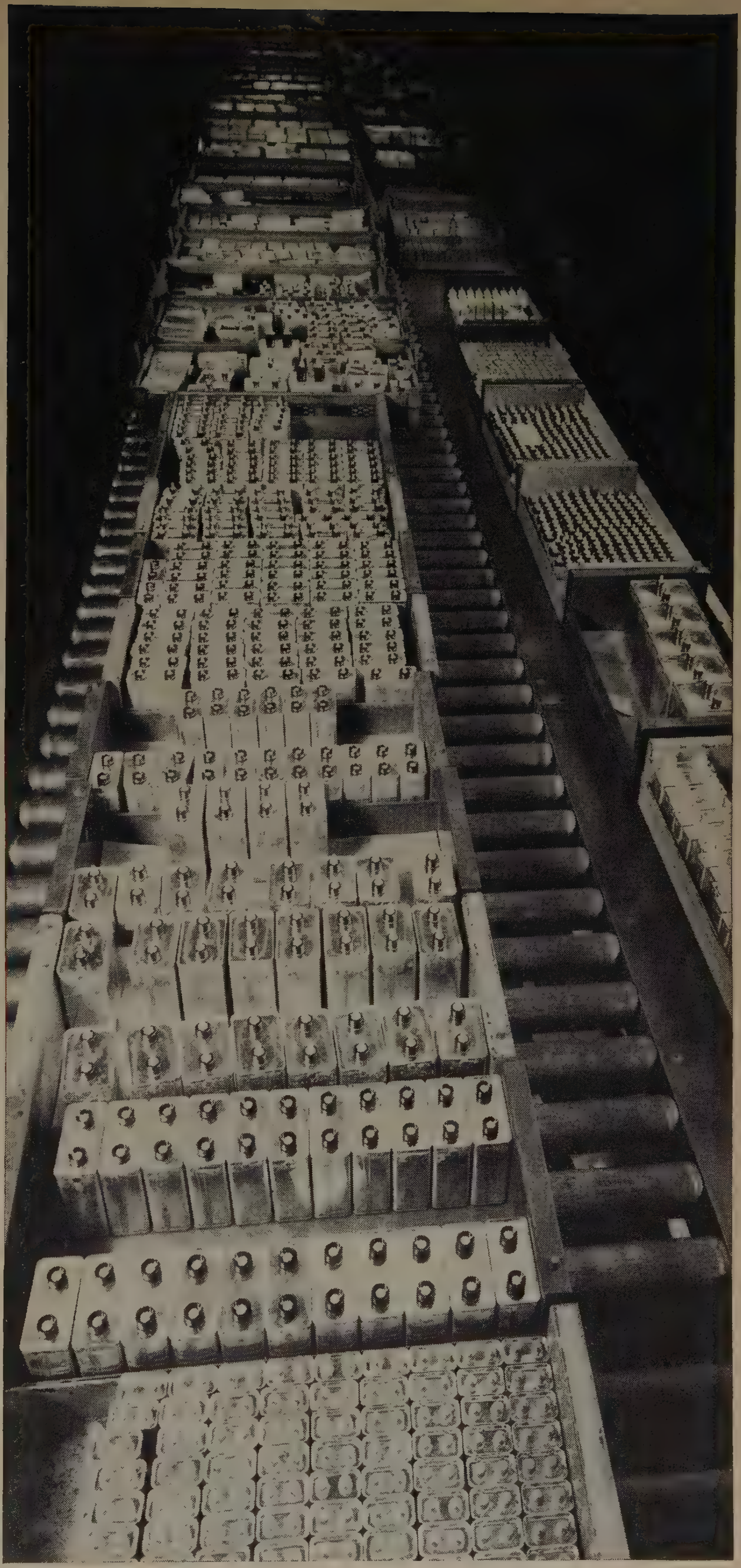

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Carlat, B. B., 108-3267 Rd., Forest Hills, N. Y.
Cohn, G. I., 265 Woodland Rd., Lake Forest, Ill.
Cole, B. M., 7710 Midfield Ave., Los Angeles 45, Calif.
Conway, J. F., Jr., 4610 Plaza Dr., Fort Wayne, Ind.
Cottony, H. V., 414 Taylor St., Chevy Chase 15, Md.
Craig, L. H., 81 Point Rd., Little Silver, N. J.
de Rosa, L. A., 192 Whittle Ave., Bloomfield, N. J.
Fathauer, G. H., 55 N. New Jersey, Indianapolis, Ind.
Fleming, L., 510 N. West St., Falls Church Va.,
Foster, C. L., 1901 West 73 Ter., Kansas City 5, Mo.
Friedberg, M. R., 2075 Revere Rd., Cleveland Heights 18, Ohio
Gerlach, G. G., R.F.D. 3, Doylestown, Pa.
Grantham, R. E., 5423 Northfield Rd., Bethesda 14, Md.
Guernsey, G. F., 5413 Huberville Rd., Dayton 3, Ohio
Hieronymus, T. G., 4209 Michigan Ave., Kansas, City 4, Mo.

Hill, B. C., Jr., Reeves-Hoffman Corp., 145 Cherry St., Carlisle, Pa.
Hoagland, C. C., 2244 Mt. Carmel Ave., Glenside, Pa.
Kreuz, L. B., Box 1624, University Station, Austin 12, Tex.
Lane, R. F., Bell Telephone Laboratories, Inc., Whippany, N. J.
Matange, N. D., 70-12 67 St., Glendale, L.I., N. Y.
Mills, M. H., 4721 N. Chelsea Lane, Bethesda 14, Md.
Moran, R. B., Jr., 170 E. Orange Grove Ave., Pasadena 3, Calif.
Oliver, G. E., 72 Orange St., Brooklyn 2, N. Y.
Press, M., 137-77 70 Ave., Kew Gardens Hills, L. I., N. Y.
Sloughter, W. J., R.F.D. 2, Ridgefield, Conn.
Smith, C. P., 904 Pleasure Rd., Lancaster, Pa.
Thompson, R. G., Columbia Broadcasting System, 485 Madison Ave., New York 22, N. Y.
Veronda, C. M., 39 Summit Rd., Port Washington, L. I., N. Y.
Weltin, O. K., 197 Lester Rd., Park Forest, Chicago Heights, Ill.
Whalley, W. B., 201-10 28 Ave., Bayside, L. I., N. Y.
Wilson, D. G., Rm. 212, Electrical Engineering Laboratory, University of Kansas, Lawrence, Kans.

## Admission to Senior Member

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Bleichner, C. W., 107 Main St., Altamont, N. Y.
Clogston, A. M., Bell Telephone Laboratories, Inc., Murray Hill, N. J.
Crawford, A. A., 456 N. Rodeo, Beverly Hills, Calif.
Dahlberg, R. S., 2505 Fernwood Ave., Roslyn, Pa.
Deschamps, G. A., 1391 Madison Ave., New York 29, N. Y.
Donath, E., 22 Madison St., Princeton, N. J.
Fox, Kenneth H., 105 Woodland Dr., Fair Haven, N. J.
Harris, J. S., 30 Havemeyer Lane, Old Greenwich, Conn.
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*(Continued on page 66A)*

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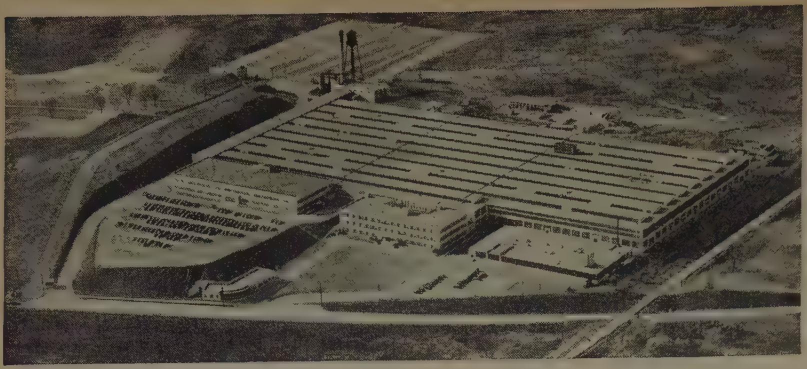

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(*Continued from page 60A*)

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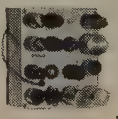

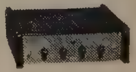

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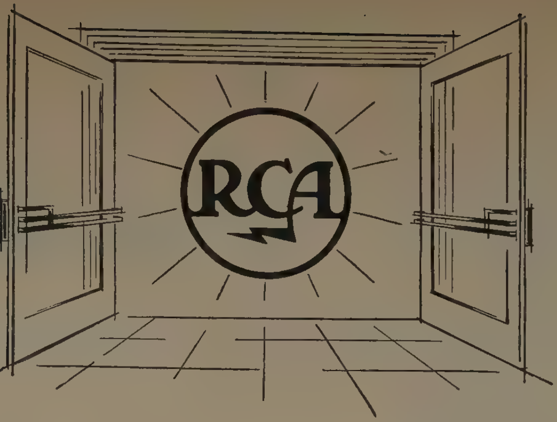

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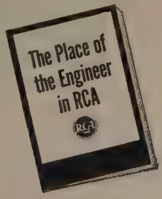
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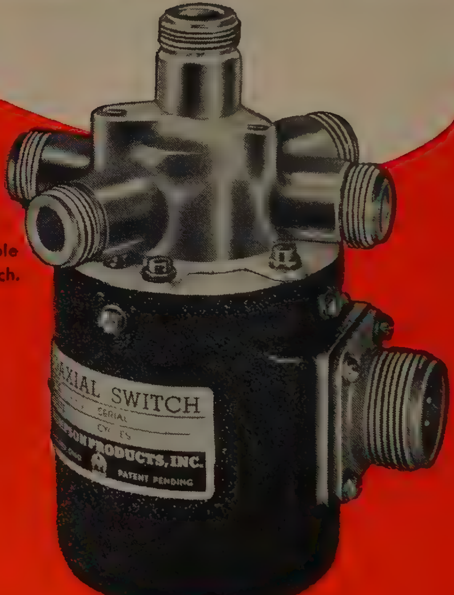
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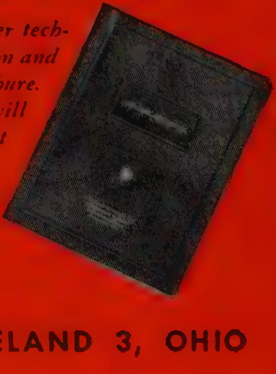

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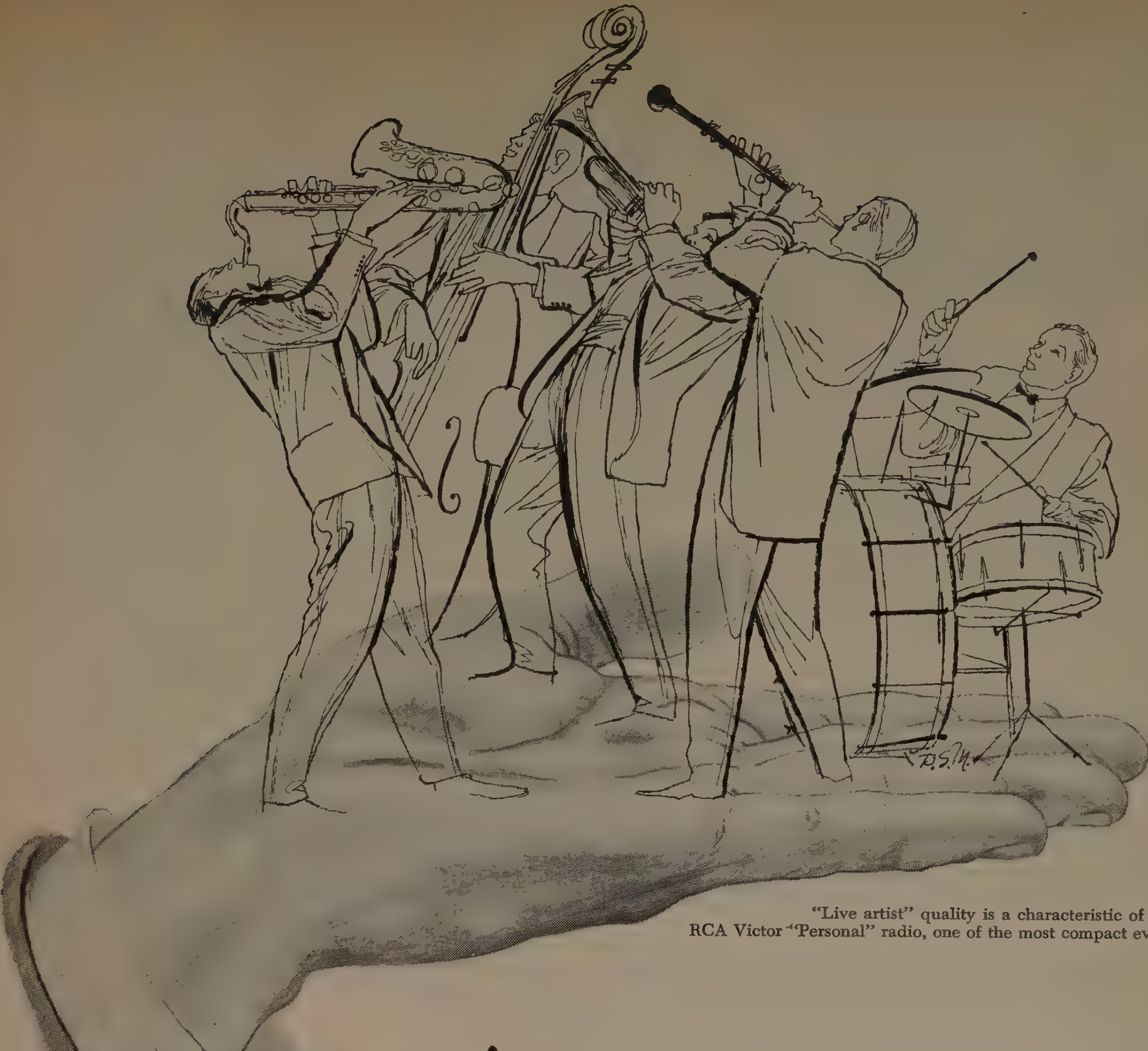

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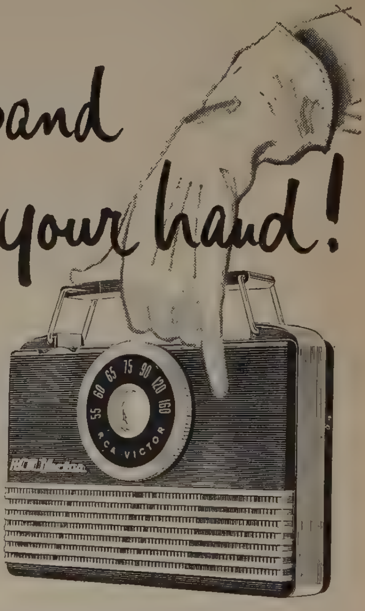

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Slaybaugh, C. W., Scarsdale Manor North, Scarsdale, N. Y.
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Middleton, W. W., Engineering Department, Bell Telephone Company of Pennsylvania, 210 Pine St., Harrisburg, Pa.
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*(Continued on page 68A)*

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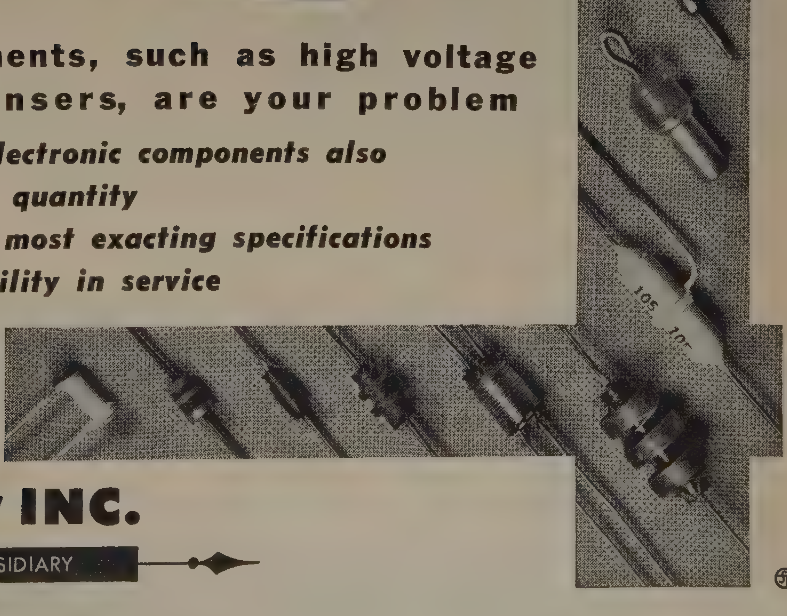

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In the frontier spectral region where radio meets infra-red, and where you can generate coherent radiation, but lack sufficient power for its electrical detection, your problem can be solved with a Golay Pneumatic Infra-red and Millimeter Waves Detector. Among the characteristics that render the Golay Detector useful for this and other purposes are:

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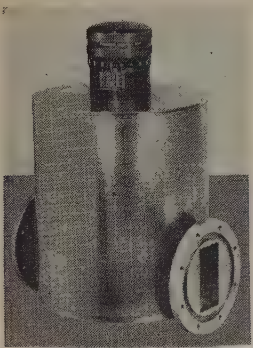

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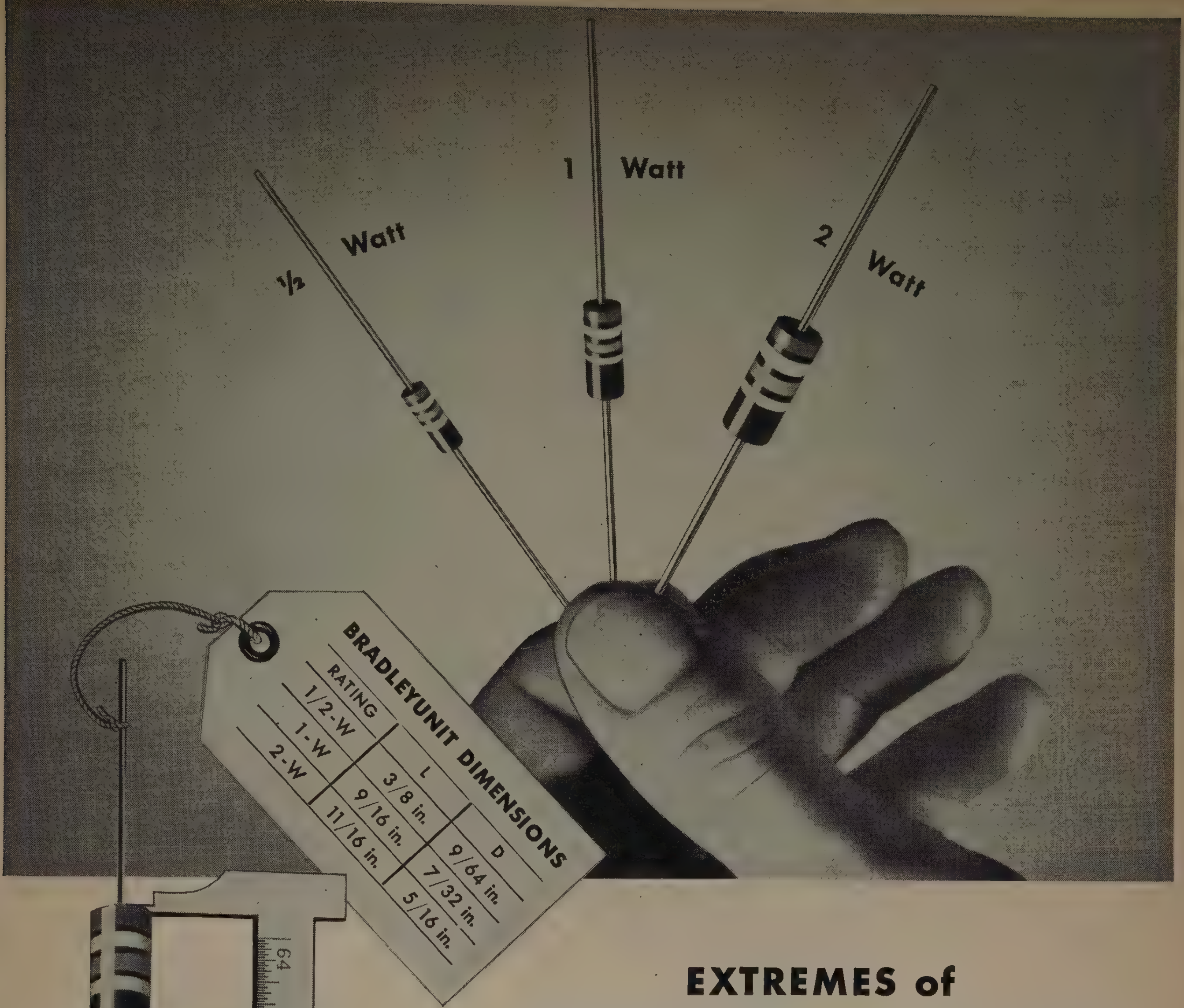

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| RATING | L | D |
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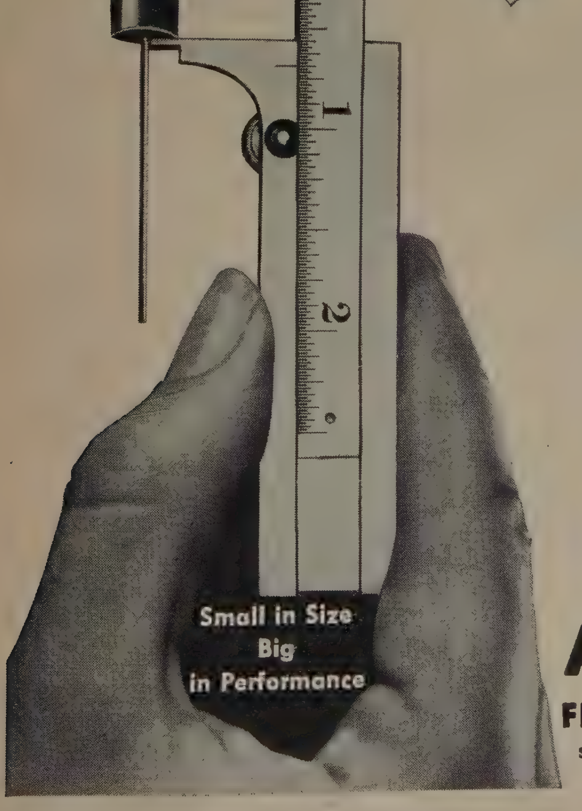
| 1/2-W | 3/8 in. | 9/64 in. |
| 1-W | 9/16 in. | 7/32 in. |
| 2-W | 11/16 in. | 5/16 in. |



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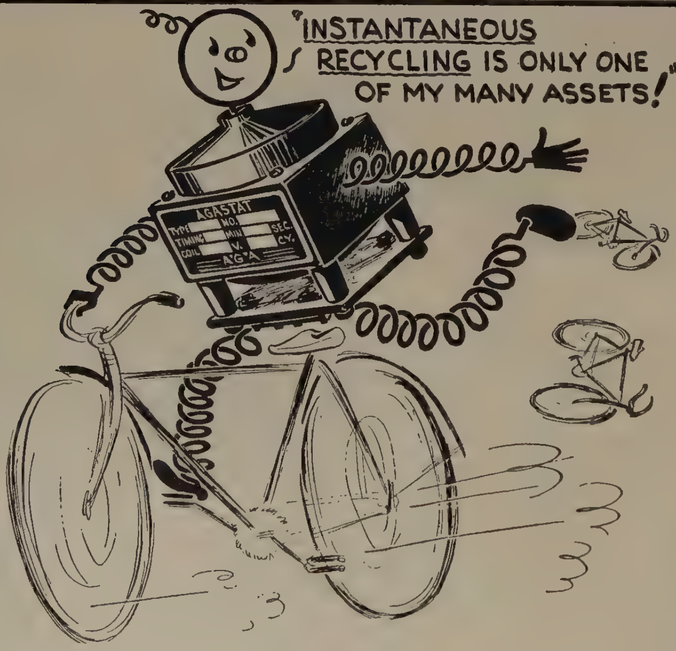
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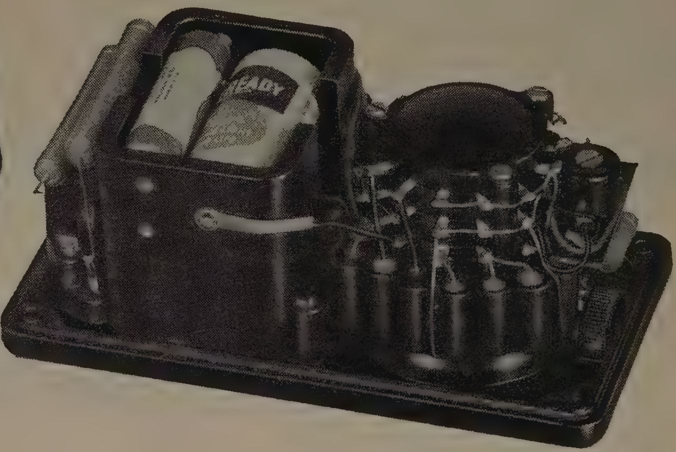
*NOTE the wide ranges of this compact pocket-size instrument. Note controls—flush with panel. Then study the inside view. Nowhere will you find, in design and manufacturing quality, the equal of 666-R.*



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### WITH SELF-CONTAINED RESISTANCE RANGES TO 3 MEGOHMS

**1. Resistance Ranges** from 0-3000 Ohms (.5 Ohm low reading) to 3 Megohms, self-contained. Also A.C.-D.C. Volts to 5000, 10 ranges; and 3 Direct Current ranges.

**2. Enclosed Selector Switch,** molded construction. Keeps dirt out, and retains contact alignment permanently.

**3. Unit Construction**—Resistors, shunts, rectifier, batteries, are housed in a molded base integral with the switch. Direct connections without cabling. No chance for shorts.

**4. Resistors** are precision film or wire-wound types, each in its own compartment.

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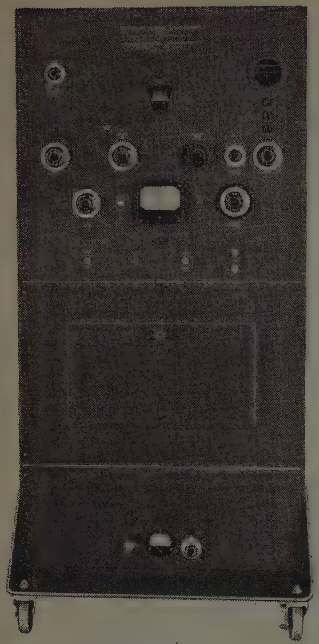
Model 90



Specifications:

CARRIER FREQUENCY

RANGE: Continuously variable from 20 to 250 megacycles, in eight ranges.

MODULATION

PERCENTAGE: Continuously variable from 0 to 100%.

ENVELOPE: Sinusoidal, or composite television.

OUTPUT

LEVEL: Continuously variable from 0.3 microvolt to 0.1 volt balanced to ground (measured at 100% modulation level).

DIMENSIONS: Height—58¾" Width—28¼" Depth—25½"

POWER SUPPLY: 117 volts, 60 cycles, 700 watts.

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The required resistor is an integral part of this assembly —"built-in."

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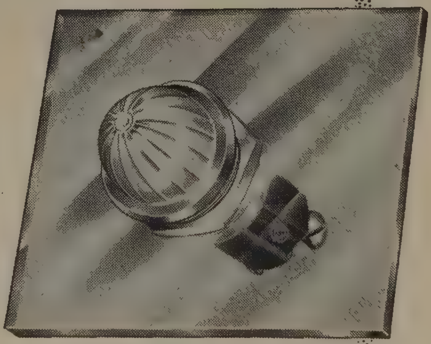

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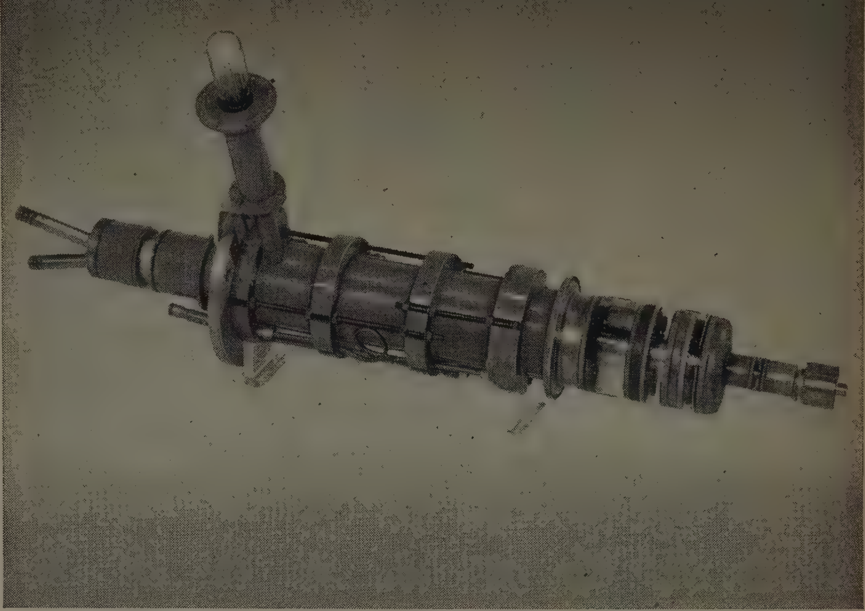
(Continued on page 76A)

VARIAN MICROWAVE ENGINEERING



... develops a new 5-kw linear-amplifier klystron

Operating in the final stage of a uhf transmitter, the new X-25 Varian Klystron provides continuous output power up to 5 kw with approximately 15-w drive. The tube is tunable from 1016 to 1056 mc and has a half-power bandwidth of about 2 mc. Gain of approximately 27 db is essentially linear to 80 per cent of maximum output.

Particularly suited to applications where crystal control and/or low-level modulation are used, the new X-25 introduces sidebands 60 db or more below the carrier and negligible noise or spurious modulation.

Long service life has been attained by use of a bombarded tantalum cathode, part of an assembly which can be replaced easily in case of accident or failure. A cascade amplifier with three cavities, the new design lends itself to stagger-tuning and other methods of broadbanding. It is typical of amplifiers practicable for other frequencies in this band.

TENTATIVE SPECIFICATIONS

**Typical Operating Characteristics**

| | |
|---|---|
| Beam Voltage, kv | 12 |
| Beam Current, amp | 1.6 |
| Power Output, max kw | 5 |
| Linear Output, max kw | 4 |
| Gain, db | 27 |
| Frequency, mc | 1016-1056 |
| Bandwidth, mc | 2 |
| Spurious Sidebands, db | —60 |

**Mechanical Characteristics**

| | |
|---|---|
| Length, overall, in. | 42 |
| Weight, approx lb | 60 |
| Input Connection | Type N |
| Output Connection | Probe to feed 4-in. by 8-in. waveguide |
| Focussing | Magnetic |
| Cooling | Water and Air |

**A counterpart of the X-25 Klystron illustrated above is the heart of the powerful new General Electric uhf television transmitter. These as well as other high-power L-band klystrons being produced by Varian Associates have a wide range of communication and industrial applications.**

VARIAN associates

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san carlos, calif.

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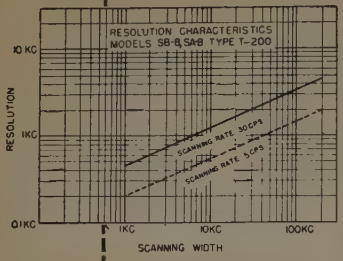

## PANORAMIC PANALYZOR SB-8 AND PANADAPTOR SA-8

These instruments enable spectrum analysis of signals so close in frequency that their corresponding indications would normally mask one another.

*FEATURES—Continuously variable resolution ● Continuously variable scanning width ● Long persistence cathode-ray tube . . . 5" Screen ● Intensity grid modulation for pulse analysis ● Synchronous and non-synchronous scanning ● Variable scanning rates ● Linear and log amplitude scales*

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*—is predicated on*

- Understanding the end use
- Electronic "know how"
- Design technique
- Dependable components

*for better instrumentation*

## Subminiature vacuum tubes

| | |
|---|---|
| 5799 | Diode |
| 5800 | Electrometer tetrode |
| 5803 | Electrometer triode |
| 5828 | Triode, medium mu |
| 5841 | Corona Regulator, 900 V |
| 5950 | Corona Regulator, 700 V |
| VXR-130 | Regulator, 130 V |
| VX-10 | Switch |

## Vibrator power supply

High conversion efficiency
●
Low weight and size

The Model 517 vibrator type power supply provides regulated high and plate voltages for counter tubes, photomultipliers and associated circuit from 4½ V of battery

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Vacuum sealed in glass in a range of values from 100 megohms to 10,000,000 megohms. A quality resistor for today's precision instrumentation.

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Square wave testing techniques come into wider use as the need for good transient response in wide band amplifiers becomes increasingly important. In order to test the high frequency response it is necessary to have a signal which has a rise and fall at least equal to and preferably faster than the risetime of the amplifier being tested. In addition to a sharp rise and fall, the test signal should be free of over-shoot and other spurious responses. For examination of the low frequency response a square wave signal having flat horizontal portions is needed.

TEKTRONIX Type 105
Square Wave Generator
Price $395.00 f.o.b.
Portland, Oregon

The TEKTRONIX **Type 105** Square Wave Generator provides a suitable signal for both of the above tests. Its frequency range, extending continuously from 25 cycles to 1 mc., combined with its risetime of .02 microseconds, makes it possible to quickly and accurately test amplifiers, filters, etc., having pass bands from a few cycles to 20 mc.

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**THE MODEL 310A measures 100 microvolts to 100 volts from 10 cycles up to 1 MC with 3% accuracy (and up to 2 MC with 5% accuracy) at *any* point on the single logarithmic voltage scale.**

- Input Impedance is 2 megohms shunted by 15 mmfds on the 0.001 and the 0.01 ranges and by 8 mmfds on the other ranges.
- Generous use of negative feedback provides customary Ballantine stability.
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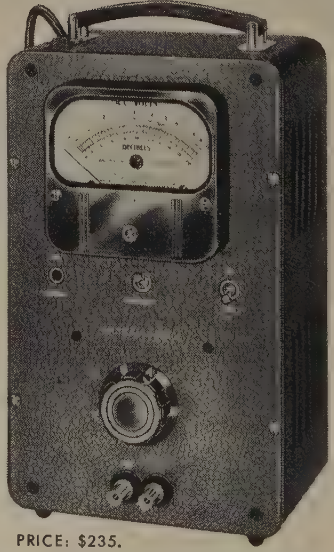
*For further information on this Voltmeter and other Ballantine Voltmeters and accessories, write for catalog.*



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*(Continued on page 79A)*

### A COMPLETE APPROACH TO

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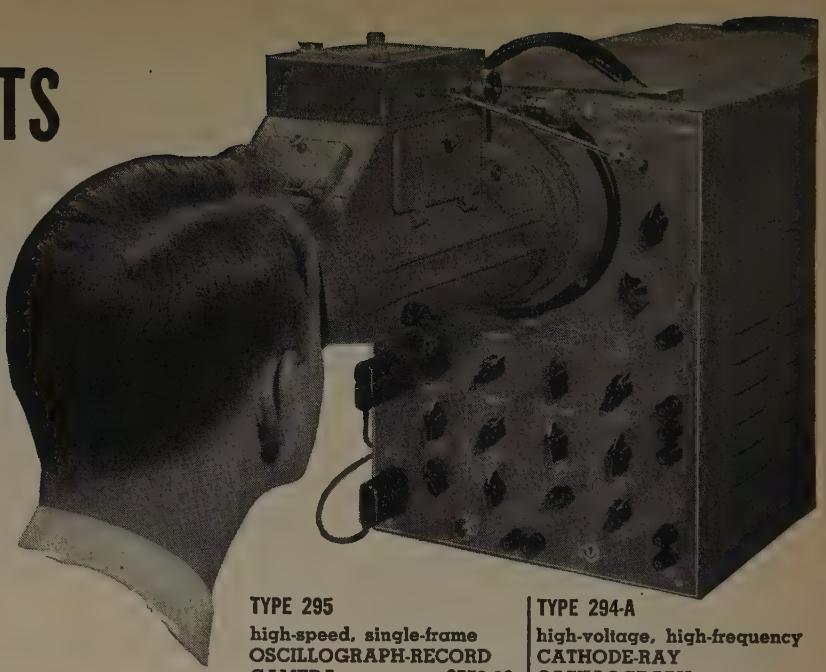
For visual observation of pulses and single transients, the Type 294-A Cathode-ray Oscillograph provides high light-output and wide-band response. For careful study and permanent reference of these signals the Type 295 Oscillograph-record Camera records writing rates as high as 35 inches per microsecond.



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high-speed, single-frame
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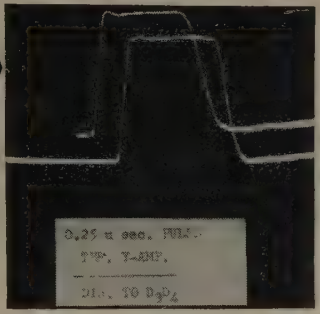
**PULSE RESPONSE**
**FREQUENCY RESPONSE**
**SENSITIVITY**

This oscillogram illustrates the double exposure technique. Binocular viewing in the Type 295 facilitates proper positioning for close comparison.



The pulses in the oscillogram at left are identical pulses of 0.25 microsecond width. The first pulse was applied through the Y-axis amplifier of the Type 294-A, and the second, directly to the vertical deflection plates. A comparison of their waveforms illustrates the excellent transient response of the Y-axis of the Type 294-A.

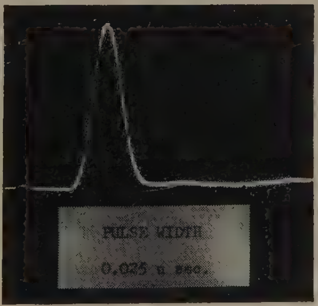
Response of the Y-axis amplifier to a rise time of 0.01 microsecond or less is 0.03 microseconds max. Notice that a minimum of overshoot (less than 2%) is introduced by the amplifier.

For the study of sinusoidal frequencies, the response of the Type 294-A extends from 10 cps. to 12 megacycles (down 30%). Sensitivity of the Y-axis, through the amplifier, is 0.42 peak-to-peak volts per inch.

---

**AVAILABLE DEFLECTION**
**LIGHT OUTPUT**
**SIGNAL DELAY**

"Time" and "Bulb" exposures may be taken with the Type 295. And provision is made so that equipment may be triggered simultaneously with shutter opening. With appropriate accelerating potentials, the Type 295 is capable of recording single transients in excess of 280 inches per microsecond.



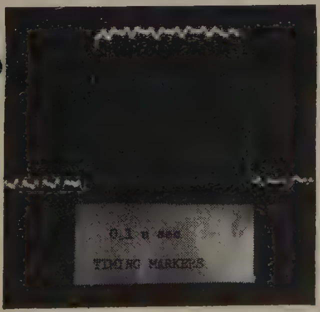
The Type 294-A provides undistorted vertical deflection of 1.3 inches or more for both positive and negative pulses; and 2.75 inches for symmetrical signals. The high light-output of the Type 294-A increases the value of the large, vertical deflection provided by the Y-axis amplifier. This is illustrated by the high visibility of the rise and decay of the pulse shown at left. Here, the Type 5XP— Cathode-ray Tube of the Type 294-A was operated at 12 kv. However, where maximum light output is not required, the accelerating potential may be lowered to 7 kv by means of a switch. At this level of operation, of course, the available undistorted deflection is increased.

To insure the complete display of fast pulses such as those at left, the Y-axis includes a 0.25 microsecond signal-delay line.

---

**SWEEP SPEEDS**
**TIME CALIBRATION**

The built-in illuminated data-card of the Type 295 will prove invaluable when making time measurements. A film take-up cassette is arranged so that exposed frames of film may be separated from unexposed film and taken to the darkroom for immediate developing.



Complementing the Y-axis performance of the Type 294-A, sweep durations are continuously variable from 0.1 second to 3 microseconds. By increasing the length of the sweep, speeds greater than 0.25 microsecond per inch may be obtained, thus providing more detail to facilitate the study of short-duration pulses.

Calibration of the sweeps of the Type 294-A is accomplished with vertical marks occurring at intervals of 100, 10, 1, or 0.1 microseconds. In the oscillogram at left, the 0.1 microsecond markers appear mixed with the signal on the vertical axis. Time measurements may also be made by double exposure of first, the signal, and second, the timing markers.

---

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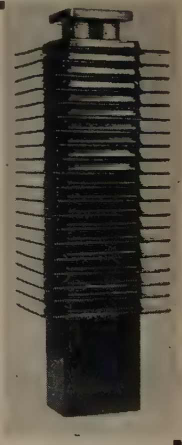

DA-21/U

Non-Standard Types *(shorter flange body than Standard)*

| Specifications | DA-21/U | DA-22/U |
|---|---|---|
| Waveguide Size (O.D.) | 5/8" x 1 1/2" | 1/2" x 1" |
| Coupling Flange | UG-51/U (plain) | UG-39/U (plain) |
| Frequency Range (mcs.) | 7,050-10,000 | 8,500-12,400 |
| V S W R | Less than 1.15 over band | |
| Power Rating: | 280 watts ave. | 175 watts ave. |
| (Forced Air Cooling) | 500 watts ave. | 300 watts ave. |
| (on 1 ms., pulses) | 0.8 megawatt peak | 0.2 megawatt peak |
| Life Rating | 100 Hours operation under continuous maximum power rating | |
| Dimensions | 11 1/2" x 2 3/4" x 3 3/4" | 9 1/2" x 2 3/4" x 3 3/4" |
| Weight | 12 lbs. | 9 lbs. |
| Unit Price | $99.50 | $89.50 |

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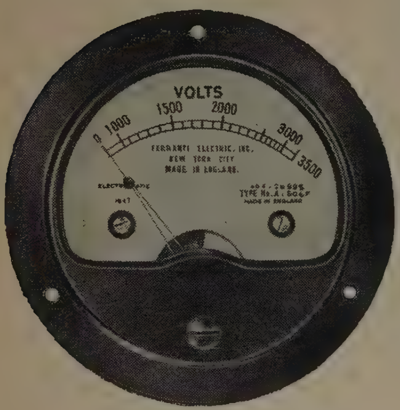

This instrument permits voltage readings on AC or DC circuits of very high resistance. The only current drawn is the very small leakage current and a very low capacitance current on AC circuits. Very useful for the many high voltage—low current circuits employed in nuclear research. Available with full scale voltages ranging between 300 and 3500 volts. Special laboratory instrument available with full scale reading of 150 volts. Full scale capacitance ranges from 8 mmfds for the 3500 volt model to 100 mmfds for the 150 volt instrument. Magnetic damping. 2½" dial. Write for complete specifications.

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JACKSON

**Vertical Sensitivity .018 RMS v.p.i.**

**Stable Band Width Thru 4.5 Mc**

# JACKSON Oscilloscope gives you "dual service"

This is a high-quality, laboratory-grade 5" Oscilloscope that provides the "dual service" of both high sensitivity and wide band width.

## specifications

**Vertical Amplifier**—Video-type frequency compensation provides flat response within 1.5 db from 20 cycles thru 4.5 Mc, dropping smoothly to a still useful value at 6 Mc.

**Sensitivity Ranges**—With a band width of 20 cycles thru 100 Kc, the sensitivity ranges are 018, 18, 1.8 RMS volts-per-inch. The wide band position 20 cycles thru 4.5 Mc has sensitivity ranges of .25, 2.5, 25 RMS volts-per-inch.

**Horizontal Amplifier**—Push-pull with sensitivity of .55 RMS volts-per-inch.

**Input Impedances**—Vertical. 1.5 megohms shunted by 20 mmfd. Direct to plates, balanced 6 megohms shunted by 11 mmfd. Horizontal· 1.1 megohms.

**Linear Sweep Oscillator**—Saw tooth wave, 20 cycles to 50 Kc in 5 steps. 60 cycle sine wave also available, as well as provision for using external sweep.

**Input Voltage Calibration**—Provides a standard voltage against which to measure voltages of signal applied to vertical input.

**Vertical Polarity Reversal**—For reversing polarity of voltage being checked or for choosing either positive or negative sync. voltages.

**Return Trace Blanking**—Electronic blanking provides clear, sharp trace to prevent confusion in waveform analysis.

**Synchronizing Input Control**—To choose among INTERNAL, EXTERNAL, 60 CYCLE, or 120 CYCLE positions.

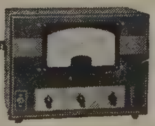
**Intensity Modulation**—60 cycle internal or provision for external voltage for intensity modulation uses.

**Additional Features**—Removable calibration screen—Accessory Model CR-P Demodulation Probe for Signal Tracing—All-steel, gray Ham-R-Tex cabinet. Total net weight only 26 pounds. Same height as other Jackson TV instruments: 13" H x 10¼" W x 15⅛" D.

**Prices:** Model CRO-2, Users' Net $197.50.
Model CR-P Probe, Users' Net $9.95.

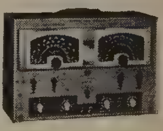
### TWO OTHER FINE JACKSON INSTRUMENTS

**Model 655 Audio Oscillator**



Sine-wave 20 cycles to 200,000 cycles. Less than 5% harmonic distortion between 30 cycles and 15,000 cycles. Frequency calibration accurate within 3% or 1 cycle. Hum level down more than 60 db of maximum power output. Output impedances of 10, 250, 500, 5000 ohms or Hi Z resistive output.

**Model TVG-2 TV Generator**



Sweep Oscillator in three ranges from 2 Mc thru 216 Mc, all on fundamentals. Reversible sweep direction. Sweep width variable .1 Mc thru 18 Mc. Marker covers 4 Mc thru 216 Mc. Crystal Oscillator to use as Marker or Calibrator. Video Modulation, from external source for using actual video signal for check, or for use with Audio Oscillator to produce bars for linearity checks.

See your electronics distributor for more information, or write

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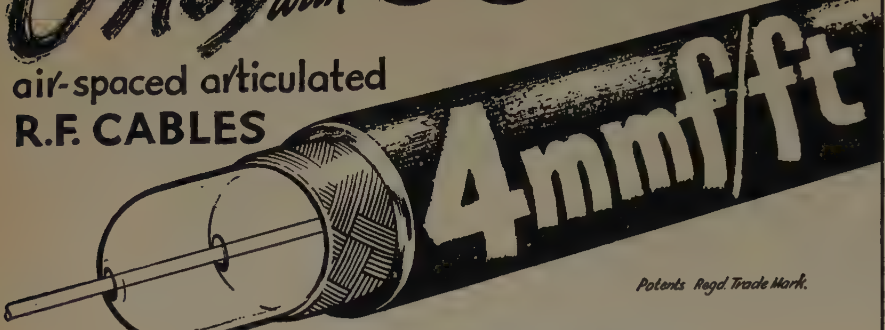

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Settlement by your own $ check*

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|---|---|---|---|---|
| A 1 | 74 | 1.7 | 0.11 | 0.36 |
| A 2 | 74 | 1.3 | 0.24 | 0.44 |
| A 34 | 73 | 0.6 | 1.5 | 0.88 |
| LOW CAPAC TYPES | CAPAC mmf/ft. | IMPED OHMS | ATTEN db/100ft 100Mc/s. | O.D." |
| C 1 | 7.3 | 150 | 2.5 | 0.36 |
| PC 1 | 10.2 | 132 | 3.1 | 0.36 |
| C 11 | 6.3 | 173 | 3.2 | 0.36 |
| C 2 | 6.3 | 171 | 2.15 | 0.44 |
| C 22 | 5.5 | 184 | 2.8 | 0.44 |
| C 3 | 5.4 | 197 | 1.9 | 0.64 |
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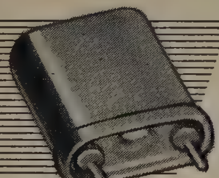
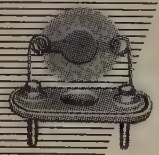
# Pacing the Field





The electronic industry is progressing with seven league strides. Probably no segment of this industry has made greater progress than that accomplished in the design and manufacture of crystals.

Stabilized, metal plated and wire mounted crystals of today are far superior in stability and precision.

Take advantage of these advances by designing your equipment to use modern crystals — such as the H-17. Compare the specifications and features of the H-17 with the old-fashioned units of 5 years ago!

**H-17 SPECIFICATIONS**—Frequency range 200 kc to 100 mc. Hermetically sealed metal holders. Wired mounted, silver plated crystals. Two type H-17 holders fit loctal socket. H-17 is military type HC6/U. CAATC #3R1-3.

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SANDWICH, ILLINOIS

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# SOLVE YOUR RADAR — CALL PAUL J. PLISHNER — SONAR SUPPLY PROBLEMS

**SIGNAL GENERATOR,** using 417A klystron, 2700-3300 mc. Output approx. 50 mw. 115 vac power supply. With tubes, new ....$425

**10 CM RF PACKAGE,** using 2J22 magnetron, freq. range 3267-3333 mc, complete with power supply and pulser giving apx. 20 kv @ 30 A. 1 usec, 1000 PPS. Power output 265 kw. ⅞" rigid coax plumbing thruout. Uses 417A klystron mixer, 6Ac7 preamp. Pulser is **715 B HARD TUBE.** Complete RF unit, pulser unit, receiver front end, new, with tubes. Requires 115v, 400 cy ac primary source ....$385

**RECEIVER POWER SUPPLY** for GL 446 type lighthouse tubes (2C40, etc.) 115 vac, 60 cycles. Panel mounting. Less tubes ....$32.50

**10 CM DISH AND DIPOLE ASSY;** apx. 30" parabola, with 360 deg. rotating mechanism, and approx. 10 deg. tilt mechanism. Operating from 24 vdc. With selsyn ....$85.00

**10 CM LOW POWER** tunable load with circ. cover .... $225.00

**COAX. CRYSTAL MOUNT,** type N connectors .$17.50

**RT-39/APG-5 10 CM** light house RF head c/o Xmtr.-Recvr-TR cavity, compl. rcvr & 30 MC IF strip using 6AK5. (2C40, 2C43, 1B27 lineup) w/Tubes.

**AN/APR5A 10 CM ANTENNA** equipment consisting of two 10 cm waveguide sections, each polarized 45 degrees ....$5.00 per set

## 3000 MC BENCH TEST PLUMBING

**POWER SPLITTER:** 726 Klystron input dual "N" output ....$5.00

**MAGNETRON COUPLING FOR TYPE 720 MAG.** to 1½" x 3" Waveguide ....$35.00

**10 CM WAVEMETER** WE type B43549 Transmission type. Type N Fittings Veeder Root Micrometer dial, Gold Plated W/Calib Chart P/o Freq. Meter X66404-A. New ....$99.50

**AS14A/AP-10 CM** Pick up Dipole with "N" Cables .... $4.50

**LHTR. LIGHTHOUSE ASSEMBLY** Part of RT39 APG 5 & APG 15. Receiver and Trans Cavities w/assoc. Tr Cavity and Type N CPLG. To Rcvr. Uses 2C40, 2C43, 1B27. Tunable APX 2400-2700 MCS. Silver Plated ....$49.50

**BEACON LIGHTHOUSE CAVITY** 10 cm with miniature 28 volt DC FM motor. Mfg. Bernard Rice ....$47.50 ea.

**MAGNETRON TO WAVEGUIDE** Coupler with 721A Duplexer Cavity, gold plated ....$45.00

**721A TR BOX** complete with tube and tuning plungers .... $12.50

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**⅞" RIGID COAXIAL TUNING STUBS** with vernier stub adjustment. Gold Plated ....$17.50

**⅞" RIGID COAX ROTARY JOINT.** Pressurized. Sperry #810613. Gold Plated ....$27.50

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**ROTARY JOINT.** Part of SCR-584 ....$35.00 ea.

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**SHORT RIGHT ANGLE BEND,** with pressurizing nipple ....$3.00

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**CG 54/U—4 foot flexible section** ¼" IC pressurized .... $15.00

**⅞ RIGID COAX,** Bead Supported ....$1.20

**SHORT RIGHT ANGLE BEND** ....$2.50

## 8500-9500 MC BENCH TEST PLUMBING

**AN/APS-15A "X"** Band compl. RF head and mod. inch 725-A mag and magnet, two 723A/B klystrons (local osc. & beacon) 1B24, TR, rcvr ampl. duplexer, HV supply blower, pulse xfmr. Peak Pwr Out: 45 KW apx. input: 115, 400 cy. Modulator pulse duration .5-2 microsec., apx. 13KV. PK. Pulse, with all tubes incl. 715B, 829B, BKR 73, two 72's. Complete pkg. ....$375.00

**COMPLETE 3 CM. RADAR SYSTEM.** 40 KW peak transmitter, pulse modulator, receiver, using 723AB power supply operating from 115V 800 Cycle, antenna system. Complete radar set neatly packaged in less than 16 cubic feet. Less receiving Type Tubes, but including all others, in used but excellent condition —$350.00. This price for laboratories, schools, and experimental purposes only.

**2J42 PULSE MODULATOR,** 14 Kw max. rating, 7kw min. Plate voltage pulsed 5.5 kv, 6.5 amp, .001 duty cycle, 2.5 usec pulse length max, filament 6.3 v, .5 amp. Includes magnetron mtg. and blower. Requires 3C45 and 2-3B24 ....$75.00

**3 CM SECTOR SCANNING ANTENNA,** 18" dish, cutler feed dipole, 24 vdc drive motor ....$75.00

**APS-3 RADAR,** new and complete, using 725A magnetron ....$950.00

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**"X" BAND PREAMPLIFIER,** consisting of 2-723 A/B local oscillator-beacon feeding waveguide and TR-ATR Duplex sect. incl. 60 pic. IF amp. ..$47.50

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**5 FT. SECTIONS** choke to cover. Silver plated ..$14.50

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**724 TR Tube** (41 TR 1) ....$2.50

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**ROTARY JOINT** choke to choke with deck mounting .... $17.50

**3 CM. MITRED ELBOW** "E" plane unplated ..$12.00

**1¼" x ⅝"** Waveguide Tunable Termination. Precision adjust ....$65.00

**LOW POWER TERMINATION** ....$25.00

**MAGIC TEE** ....$45.00

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**WAVEGUIDE LENGTHS.** Cut to size and supplied with 1 choke, 1 cover, per length ....$2.00 per ft.

**BI DIR-COUPLER** WG output calibrated —25 db nominal ....$17.50

**FLEX SECTIONS,** 12" Rubber Coated ....$14.50

**MITRED ELBOW** H Plane UG51-UG52 ....$12.00

**6" ST. SECT.** choke to choke ....$3.50

**APQ 13** Constant Z Rotat Jnt. ....$22.50

**CG 98B/APQ 13** 12" Flex. Sect. 1¼" x ⅝" OD .... $10.00

**WAVE GD. RUN 1¼" x ⅝"** Gd. consists of 4 ft. sect. w/RT angle bend on one end, 2" 45 deg bend on other end ....$8.00

**X BAND WAVE GD.** 1¼" x ⅝" O.D. 1/16" wall aluminum ....Per ft. 75¢

**SLUG. TUNER ATTENUATOR** W.E. guide. Gold Plated ....$6.50

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**WAVEMETER,** 8500 to 9400 Mcs., with calibration. Micrometer adjust head. Reaction type ....$85.00

**90 DEGREES ELBOWS.** E or H plane, 2½" radius .... $12.50

**90 DEGREE TWIST.** 6" long ....$8.00

**BULKHEAD FEED-THRU ASSEMBLY** ....$15.00

**PRESSURE GAUGE SECTION** 15 lb. gauge and press nipple ....$10.00

**PRESSURE GAUGE.** 15 lbs. ....$2.50

**DUAL OSCILLATOR-BEACON MOUNT.** P/O APS10 Radar for mounting two 723A/B klystron with crystal mts. matching slugs, shields ....$42.50

**DUAL OSCILLATOR, MOUNT.** (Back to back) with crystal mount, tunable termination attenuating slugs ....$18.50

**DIRECTIONAL COUPLER.** UG-40/U Take off 20 DB .... $17.50

**ROTARY JOINT.** Choke to Choke ....$10.00

**2K25/723 AB RECEIVER** local oscillator Klystron Mount. complete with crystal mount. Iris coupling and choke coupling to TR ....$22.50

**TR-ATR DUPLEXER** section for above ....$8.50

**CU 105/APS 31** Directional Coupler 25 DB ....$25.00

**723AB MIXER**—Beacon dual Osc. Mnt. w/xtal holder .... $12.00

**TR-ATR SECT** APS 15 for 1B24 w/724 ATR cavity w/1B24 & 724 Tubes. Complete ....$21.00

**STABILIZER CAVITY** with bellows ....$21.50

**3 CM 180° BEND,** with pressurizing nipple ..ea. $6.00

**3 CM. 90° BEND,** 14" long 90° twist with pressurizing nipple ....ea. $6.00

**3 CM. "S" CURVE** 18" long ....ea. $5.50

**3 CM. "S" CURVE** 6" long ....ea. $3.50

**3 CM. RIGHT ANGLE BENDS.** "E" plane 18" long cover to cover ....ea. $6.50

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**3 CM. DIRECTIONAL COUPLER.** One way waveguide output ....ea. $15.00

**CIRCULAR CHOKE FLANGES,** solid brass .....55¢

**SQ. FLANGES,** Flat Brass ....ea. 55¢

## 6000 Mc to 8500 Mc BENCH TEST PLUMBING

### 1½" x ¾" Waveguide

**KLYSTRON MOUNT.** DB356 complete with shield and tunable termination ....$125.00

**FLAP ATTENUATOR.** DB361 ....$45.00

**VARIABLE STUB TUNER** ....$90.00

**WVGD. TO TYPE "N" ADAPTER** ....$18.50

**WAVEMETER TEE.** DB352 ....$32.50

**MAGIC TEE** ....$80.00

**DIRECTIONAL COUPLER,** two hole 25DB coupling, type "N" output ....$25.00

**PRECISION CRYSTAL MOUNT.** Equipped with tuning slugs and tunable termination ....$125.00

**TUNABLE TERMINATION.** Precision adjust ..$70.00

**LOW POWER LOAD** ....$35.00

## 4000 to 6000 Mcs BENCH TEST PLUMBING

### 2" x 1" Waveguide

**FLAP ATTENUATOR** ....$48.00

**VARIABLE STUB TUNER AND LOW POWER TERMINATION** ....$48.00

**WAVEMETER TEE** ....$48.00

**ADAPTERS:**
- Choke to choke ....$18.00
- Cover to cover ....$14.00
- Choke to cover ....$16.00

**WAVEGUIDE TO TYPE "N" ADAPTER** ....$45.00

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**CRYSTAL MOUNT.** Equipped with tunable termination and micrometer adjust crystal tuning ....$125.00

**TUNABLE TERMINATION.** Precision adjust ...$90.00

## WAVEGUIDE

⅝" x 1¼" OD ....$1.65 per foot
⅝" x 1¼" OD Aluminum .... .75 per foot
1½" x 3" OD ....$3.00 per foot
2½" x 3" OD ....$3.50 per foot
1" x ½ OD Flexible ....$4.00 per foot
⅞" rigid coax ¼" IC ....$1.20 per foot
(Available in 10 ft. to 15 ft. lengths or smaller ....$8.50 each

## 23,000 to 27,000 Mc BENCH TEST PLUMBING

### ½" to ¼" Waveguide

**LOW POWER LOAD** ....$20.00

**SHUNT TEE** ....$35.00

**WAVEGUIDE LENGTHS** 2" to 6" long, gold plated with circular flanges and coupling nuts ..$2.25 per inch

**APS-34 ROTATING JOINT** ....$49.50

**RIGHT ANGLE BEND** E or H Plane, specify combination of couplings desired ....$12.00

**45° BEND** E or H Plane. Choke to cover ....$12.00

**MITERED ELBOW,** cover to cover ....$4.00

**TR-ATR-SECTION.** Choke to cover ....$4.00

**FLEXIBLE SECTION** 1" choke to choke ....$5.00

**"S" CURVE CHOKE** to cover ....$4.50

**ADAPTER,** round to square cover ....$5.00

**FEEDBACK** to Parabola Horn with pressurized window ....$27.50

**90° TWIST** ....$10.00

**"K" BAND DIRECTIONAL COUPLER** ....$49.50 ea.

## PULSE EQUIPMENT

**MIT. MOD. 3 HARD TUBE PULSER:** Output Pulse Power 144 KW (12 KV at 12 Amp). Duty Ratio: 001 max. Pulse duration: 5. 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses: 1-715B, 4-829-B, 3-'72's, 1-'73. New ....$110.00

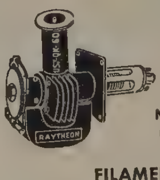
**APQ-13 PULSE MODULATOR.** Pulse Width .5 to 1.1 Micro Sec Rep, rate 624 to 1348 Pps. Pk pwr. out 35 KW Energy 0.018 Joules ....$49.00

**TPS-3 PULSE MODULATOR.** Pk. power 50 amp. 24 KW (1200 KW pk); pulse rate 200 PPS. 1.5 microsec. pulse line inpedance 50 ohms. Circuit—series charging version of DC Resonance type. Uses two 705-A's as rectifiers. 115 v. 400 cycle input. New with all tubes ....$49.50

**APS-10 MODULATOR DECK.** Complete, less tubes ....$75.00

## MAGNETRONS

| Tube | Frq. Range | Pk. Pwt. Output | Price |
|---|---|---|---|
| 2J31 | 2820-2860 mc. | 265 KW | $38.50 |
| 2J21-A | 9345-9405 mc. | 50 KW | $12.50 |
| 2J22 | 3267-3333 mc. | 265 KW | $25.00 |
| 2J26 | 2992-3019 mc. | 275 KW | $37.50 |
| 2J27 | 2965-2992 mc. | 275 KW | $25.00 |
| 2J32 | 2780-2820 mc. | 285 KW | $45.00 |
| 2J38 Pkg. | 3249-3263 mc. | 5 KW | $39.50 |
| 2J39 Pkg. | 3267-3333 mc. | 87 KW | $39.50 |
| 2J49 | 9000-9160 mc. | 58 KW | $75.00 |
| 2J61 | 3000-3100 mc. | 35 KW | $75.00 |
| 2J62 | 2914-3010 mc. | 35 KW | $75.00 |
| 3J31 | 24,000 mc. | 50 KW | $85.00 |
| 718DY | 2720-2890 mc. | 250 KW | $35.00 |
| 720BY | 2800 mc. | 1000 KW | $75.00 |
| 720CY | 2860 mc. | 1000 KW | $75.00 |
| 725-A | 9345-9405 mc. | 50 KW | $50.00 |
| 730-A | 9345-9405 mc. | 50 KW | $50.00 |
| 700 | A, B, C, D |  | $35.00 |
| 706 | AY, BY, DY, EY, FY, GY |  | $55.00 |
| Klystrons | 723A/B $18.95, 707B W/Cavity |  | $22.50 |



### CW MAGNETRONS

| | | |
|---|---|---|
| QK 62 | 3150-3375 | mc. |
| QK 59 | 2675-2900 | mc. |
| QK 61 | 2975-3200 | mc. |
| QK 60 | 2800-3025 | mc. |

New, Guaranteed ...Each $65.00
QK 915 ....$150.00

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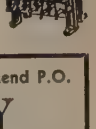
for above 115V/60 cy Pri: four 6.3V/4A Sec. 5000VT ....$27.50



Magnetron Kit of four QK's 2675-3375 inc. w/trans special ...$250.00

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Because they combine exceptional *resiliency* with good *conductivity*, Metex Electronic Products made of knitted wire mesh offer an unusually effective means of sealing and shielding a wide variety of types of electronic equipment.

As closures for sheet metal cabinets, for instance, the resiliency of "Metex" assures positive conductive contact *at every point* between cover and cabinet—eliminating costly machining for close tolerances. Metex Gaskets also assure resilient metal-to-metal contact between flanges, and can be combined with rubber compounds to function both as a shield and as a seal.

Applications in which "Electronic Weather Strips" have already proved their effectiveness include pulse modulator shields, wave-guide choke-flange gaskets, replacement of beryllium-copper fingers and springs on TR and ATR tubes.

We will be glad to put our experience at your disposal. A letter to Mr. R. L. Hartwell, outlining your problem, will receive immediate study. For preliminary information, write for bulletin "Metex 'Electronic Weather Strips.'"

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METAL TEXTILE CORPORATION

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STABLE • DEPENDABLE • MODERATELY PRICED

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- OUTPUT #1: 200 to 325 Volts DC at 100 ma regulated
- OUTPUT #2: 6.3 Volts AC CT at 3A unregulated
- RIPPLE OUTPUT: Less than 10 millivolts rms

WIDTH 14"
DEPTH 6"
HEIGHT 8"
WT: 17 LBS.

For complete information write for Bulletin G

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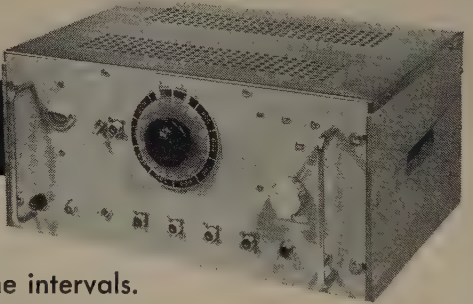
CORONA NEW YORK



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- A precision device for the generation of accurate and variable time intervals.

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MODEL 410-A

ULTRA-LOW FREQUENCY OSCILLATOR

with frequency range of 0.02-20,000 cps

FEATURING:

1. BOTH SINE AND SQUARE WAVE. 2. COMPLETELY ELECTRONIC OPERATION. 3. EXCELLENT AMPLITUDE CONSTANCY. 4. LOW DRIFT AND DISTORTION.

*DESCRIPTION:* This oscillator, Model 410-A, covers the sub-audio and the entire audio range. It provides both sine and square wave at any frequency range between 0.02 and 20,000 cps.

Precisely engineered and constructed, the Model 410-A is ideal for medical research, geophysical and seismological instruments, and design and development of servo-mechanisms and vibration controls.

SPECIFICATIONS:

1. FREQUENCY RANGE: 0.02 to 20,000 cps, continuously variable in six decade bands.
2. SINE WAVE AMPLITUDE: Varies less than ± 0.25 db (3%) over the entire frequency range.
3. DISTORTION & HUM: Less than 0.25% over the entire frequency range.

K-H KROHN-HITE INSTRUMENT CO.
580 MASSACHUSETTS AVE., CAMBRIDGE 39, MASS., U.S.A.

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(Continued on page 104A)

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CABLES { TWISTED, PARALLEL, BRAIDED

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105 YEARS

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***You'll find the details on the inside front cover next month . . . be sure to see them.***

AMPEREX ELECTRONIC CORP.
25 WASHINGTON STREET, BROOKLYN 1, N. Y.
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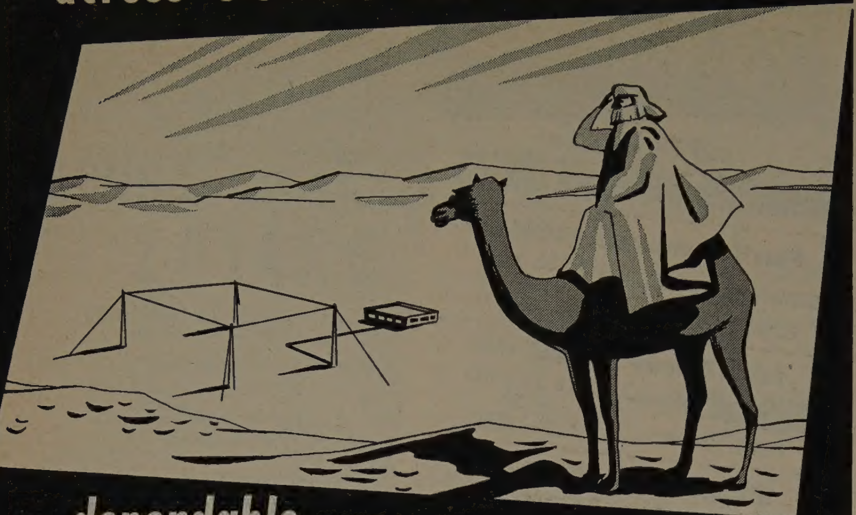
AMPEREX

for Uninterrupted Communications
across OCEANS and CONTINENTS...



dependable
ANDREW Rhombic Antenna Equipment

**Whether your problem is uninterrupted communication half-way around the world . . . or only 100 miles . . . ANDREW offers you (1) a world-wide reputation of reliability and (2) the convenience of obtaining all necessary equipment from one dependable source.**

- **Receiver Coupling Unit** efficiently distributes the output of one antenna among as many as 10 receivers. Interaction between receivers is held to negligible levels. Power gain is approximately unity (0 db) over the entire range of operation. A 4-channel unit is also available.
- **Rhombic Receiving Antenna Kit** contains in one "package" everything you need for an antenna except poles.
- **Transmitting antennas** available on special order.
- **Rhombic Antenna Coupling Transformer** is a broad band, low loss unit which matches the balanced impedance of the rhombic to the unbalanced impedance of a coaxial line.
- **Transmitting Rhombic Tuning Units** for single or multiple frequencies are available on special order.

**For Rapid, Frequent Changes in COAXIAL CIRCUITS...**

- **(a) Coaxial Patch Panel** has 24 jacks. Fits 19" relay rack. Facilitates switching coaxial circuits.
- **(b) ANDREW Coaxial Jacks and Plugs** are simple to install. No soldering through a window. Just remove one screw, slide the sections apart and solder.

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ANTENNA TUNING UNITS · TOWER LIGHTING EQUIPMENT

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look
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is
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## "GREY TIGER"

### Paper Tubular Capacitors

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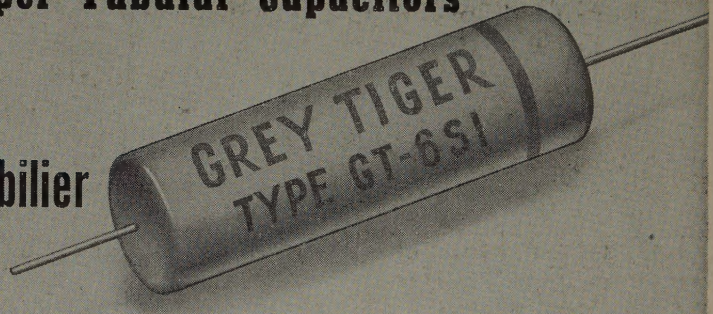

1. "Vikane"* impregnation — an exclusive C-D development—results in extra long life and outstanding performance over a temperature range of —35°C. to 100°C. Special high temperature and moisture end-fill seals unit.
2. A tubular of exceptionally high insulation resistance. At 25°C., resistance is over 10,000 megs per unit or 2,000 megs per mfd.
3. A standardized line of paper tubulars that will fill both high and low temperature requirements. No duplication of stock and record keeping! No production delays! Less inventory loss! A feature of "Vikane" impregnation.
4. A tubular with a power-factor average of 0.35% at 1,000 cycles.
5. A tubular with excellent capacity stability over both the high, low temperature range.

Tinned-copper leads, standard sizes. Attractive grey case, red lettering. Available in all commercial capacity and voltage ratings.

For further details on the "Grey Tiger" line of paper tubulars, write for Bulletin No. NB116. CORNELL-DUBILIER ELECTRIC CORPORATION, Dept. M-61 South Plainfield, New Jersey. Other plants in New Bedford, Brookline and Worcester, Mass.; Providence, R. I., Indianapolis, Ind., and subsidiary, The Radiart Corp., Cleveland, Ohio.

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# A-C Vacuum-Tube Voltmeter

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GENERAL RADIO COMPANY
VACUUM-TUBE VOLTMETER
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RANGE
OFF 1.5 5 15 50 150
ZERO
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**Five Ranges** *Four Scales cover the ranges from 0.1 to 150 volts a-c (full scale 1.5, 5, 15, 50, and 150 volts).*

**Accurate** *± 3% of full scale on all ranges; r-m-s values of sine-wave voltage.*

**Frequency Range** - *Without correction, up to 120Mc with maximum error of 10%. Correction curve supplied.*

**Input Impedance** *Equivalent input capacitance of probe is 11 μμf; with plug connectors 12 μμf; equivalent parallel input resistance 7.7 megohms at low frequencies.*

**Single Zero Adjustment** *for all ranges.*

**Internal Calibration Control** *Single adjustable resistor corrects calibration if amplifier tube is changed.*

**Auxiliary Connectors** *a G-R double plug, pair of 30" test leads, pair of test prods and two alligator clips supplied as accessories.*

**Small—Light** *only 9¼ lbs.*

**Type 1803-A**
**Vacuum-Tube Voltmeter**
**$155.**

**The Probe** *with its completely shielded case removed. Twin diode tube in probe has an inactive section connected to the grid of one triode in V-2 amplifier while active section is connected to the grid of the other triode, both sections of the amplifier being used in a balanced circuit. The balanced amplifier insures very little zero shift when line voltage varies.*

THROUGH the elimination of many unnecessary frills and extra circuit refinements which would be necessary in a meter with ohmeter and d-c circuits and scales, G-R presents this a-c vacuum-tube voltmeter with a straightforward circuit and with accuracies sufficient for most laboratory requirements at a very moderate price.

Substantially duplicating the performance of the very popular pre-war Type 726-A instrument, this new Type 1803-A Vacuum-Tube Voltmeter sells for less than its predecessor and is improved over the older model in that it is smaller, lighter, has a probe that is also smaller and completely shielded, has a single zero adjustment for all ranges and a power supply not limited to operation at a single frequency.

The probe plugs into clips on the side of the cabinet, in which position the auxiliary test leads and terminals supplied with the instrument can be conveniently attached to the input connections.

***Write for Complete Data***

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